GORDON J. KING

DUAL-STANDARD AND 625-LINE TELEVISION RECEIVERS

NORMAN PRICE

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Assoc.I.E.R.E., M.T.S.

This book has been written for the service technician and enthusiast and it will bring them up to date with the latest television developments.

Commencing with the u.h.f. tuner, the book proceeds to detail the circuit alterations and arrangements in the vision signal, audio signal, synchronizing, contrast control, a.g.c. and time-base circuits which are necessary for the correct display of 625-line pictures, and which must be switched from one standard to the other in dual-standard receivers.

U.H.F. signal propagation, and the important features of suitable aerials are dealt with, and there is a useful chapter on servicing techniques for the new standard.

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AUTHOR'S PREFACE

THIS is not a highly technical book. It has been written with the service technician and enthusiast fully in mind. The take-off level assumes that the reader has a reasonable knowledge of the basic concepts of television transmission and reception.

In a book of this size it is not possible, of course, to explore avenues of basic television principles—many excellent books are already available on this subject. However, it has been my aim to focus attention on specific principles where knowledge of these is particularly needed for the understanding of a dual-standard or 625-line technique.

Dual-standard sets will be with us for many years to come. Even when 405-line transmissions eventually cease (no one knows yet when this will be) and 625-line-only models are in the shops, dual-standard sets will continue in use for many years afterwards—just as 405-line-only sets of ten-year or more vintage are still being used in Great Britain today.

The book investigates the 405-line and 625-line signal characteristics and compares one with the other. It goes on to analyse in some detail 625-line and dual-standard circuits as they apply to the very latest models.

Information on u.h.f. channels is also given and I have devoted an entire chapter to the important subjects of u.h.f. propagation and aerials. There are also chapters on conversion and servicing.

I should like to convey my sincere thanks to the many manufacturers and to my friends in the television industry who have helped to make this book possible. My thanks also go to those manufacturers who have freely supplied photographs, circuit details and information on dual-standard conversion procedures; also to Mr. Read and Pye Limited of Cambridge for their assistance in making available a test receiver.

I duly acknowledge extracts from some of my articles which have appeared in *Electrical and Radio Trading*, and edited by Roy Norris.

Finally, it is hoped that this book will fill a gap which, at the time of writing, exists in the literature for the television technician and enthusiast.

GORDON J. KING

Brixham, Devon, 1963

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I

INTRODUCTION

A DUAL-STANDARD television receiver is one which works equally well on both the original British 405-line standard and the recently introduced British 625-line standard, the standard change being accomplished by a user's main control knob, sometimes coupled to the v.h.f. tuner channel selector knob.

While dual-standard operation is new to Great Britain, multi-standard operation has been practised for a number of years in European countries. For instance, it is only recently in Belgium that four-standard models have been replaced by five-standard sets!

It is unlikely that Great Britain will see the launching of receivers other than those for dual-standard working, though experimenters may investigate the possibilities of adapting dual-standard models or British 625-line-only models (when these eventually become available) for DX working on the European 625-line standards. These differ essentially from the British 625-line standard in that the separation between the sound and vision carriers is mainly 5·5Mc/s (11·15Mc/s and 6·5Mc/s French stations) as compared with the 6Mc/s separation of the British system.

The British 625-line system is to be geared to provide country-wide second programmes of the BBC and ITA, whilst retaining the 405-line system for the time being to accommodate the first programmes of these authorities.

Country-wide coverage of the BBC and ITA 405-line programmes has used up all the very high-frequency (v.h.f.) channels in Band I (41 to 68Mc/s) and Band III (174 to 261Mc/s) thereby making it necessary to enter the ultra high-frequency spectrum in Band IV (470 to 585Mc/s) and Band V (610 to 960Mc/s) for country-wide coverage of the two second programmes, and for the eventual accommodation of additional programmes. The v.h.f. channels are numbered 1 to 5 (Band I) and 6 to 13 (Band III) and the u.h.f. channels 21 to 34 (Band IV) and 39 to 68 (Band V). The London area BBC2 transmitter uses Channel 33.

U.H.F. Channels

Each area station will have four transmitters, taking up four u.h.f. channels, but not adjacent channels. In most areas a spectrum 88Mc/s wide will be put over to each station, and since each u.h.f. channel is

8Mc/s wide (see Fig. 1.1) a piece of spectrum equal in width to eleven channels is tied to each area station.

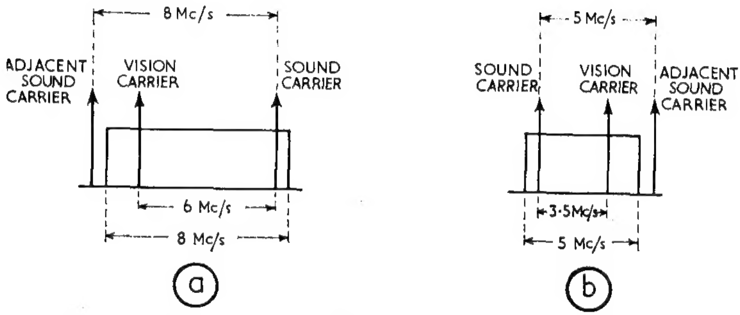


FIG. 1.1.(a) BRITISH 625-LINE CHANNEL; AND (b) 405-CHANNEL. NOTE HOW THE RELATIVE POSITION OF THE SOUND CARRIER IS CHANGED, ALSO THE FREQUENCY SEPARATION BETWEEN THE SOUND AND VISION CARRIERS

Each of the four area channels is identified relative to the lowest frequency channel of the group. The lowest frequency channel is signified by the letter "n", and the London area channels, for example, are n, n+3, n+7 and n+10. Here n is channel 23, which means that the four London channels are 23, 26, 30 and 33. The number indicates the number of channels up from the lowest frequency channel. Thus, if n is Channel 21, then n+4 is, of course, channel 25.

This method of identification allows the frequency of any channel of a group to be determined provided the frequency of the lowest number channel is known. The vision carrier of Channel 23 (London), for example, is 487·25Mc/s. Thus, n+3 vision carrier frequency is equal to 487·25 + (3×8) Mc/s, that is 511·25Mc/s, corresponding to Channel 26. Clearly, the frequency of n+3 is three 8Mc/s channel widths up from the frequency of n.

The London channel group is non-standard, and there will be other non-standard groups in some areas. However, the "standard" lattice of n, n+3, n+6 and n+10 or n, n+4, n+7 and n+10 will be adopted in the majority of areas. The present planning is geared to a maximum of six national television programmes: two on the v.h.f. bands; and four on the u.h.f. bands.

It is possible that the 405-line system will ultimately be abandoned, and it is likely that during the 625-line changeover the first programmes of the BBC and ITA will be radiated simultaneously at 405 lines v.h.f. and 625 lines u.h.f., thereby fully accommodating the four u.h.f. group channels until the first programmes are put back again into the v.h.f. channels, this time on 625 lines. The situation at that time would then be a total of four programmes all on 625 lines: two on the v.h.f. bands and two on the u.h.f. bands, leaving the stage set for the introduction of two extra programmes on the u.h.f. bands—but that is well into the future. It is important to remember that 405-line-only sets are not immediately rendered obsolete by the introduction of the 625-line system. It simply means that while the old models can only receive one or two programmes

the new dual-standard models can receive three programmes, the extra BBC2, and later a fourth, being ITV2.

Definition

We must now consider how the 405-line signals differ from those of the new 625-line system. Firstly, there is the consideration of definition. The definition of a television picture is related to the number of lines upon which it is built and **also** upon how quickly the scanning spot can change brightness as it is deflected horizontally across the screen.

The number of lines determines the **vertical** definition of a picture, the greater the number the better the definition. From this aspect, therefore, the change from 405 to 625 lines is one towards improved definition: but that is not the whole of the story!

The field frequency is locked to the 50c/s power frequency on 405 lines (so as to avoid the rippling effect which can result from asynchronous working due to the residual hum in the vision stages of the set being slightly out of step with the field frequency) and is almost at the same frequency on 625 lines. This means that on 625 lines the scanning spot needs to traverse from the left to the right of the screen on each scan in approximately two-thirds of the time it takes to scan a 405 line. This is necessary to accommodate the greater number of lines during the 50c/s field frequency period.

The maximum rate at which the spot can change brightness is governed by the bandwidth of the video channel as a whole, and it is this which determines the **horizontal** definition of a picture. On 405 lines the video bandwidth is in the order of 2·7Mc/s, which provides a horizontal definition to match that of the vertical.

The overall definition of a picture is thus dependent on **both** the vertical and the horizontal definition. Simply stepping up the number of lines without increasing the video bandwidth would not improve the **overall** definition. Indeed, it might appear to worsen it.

One way to understand this matter of definition is to consider what is required to resolve the most detailed kind of picture—a checker-board pattern of black and white squares.

Ignoring time lost during the line and field flybacks, the scanning spot must sweep all the lines in 1/25th second (remembering that there are half the total number of lines during each field and that two fields each lasting 1/50th second are interlaced—on both standards—to produce a complete picture), alternating from white to black with each two squares.

So with 405 lines and equal horizontal and vertical definition (and, for simplicity, assuming a square picture), the upper video frequency required for modulating the scanning spot is approximately $400 \times 400 \times 25$ divided by 2 (since there are two checker-board squares per cycle), or 2Mc/s.

In point of fact, because the picture is wider than it is deep and because of "lost" flyback time, the frequency is higher. As already indicated, the theoretical 405-line bandwidth extends to about 2·7Mc/s at half-power (3dB) points.

Now it is clear from the above calculation that if the horizontal and vertical definitions are **not** equal, an increase in one must be balanced by a decrease in the other if the bandwidth is to remain the same.

For instance, if 625 lines were used without any change in bandwidth the horizontal definition would have to come down to 266 approximately. Check: $600 \times 266 \times 25$ divided by 2 equals 2Mc/s (compare with the previous calculation along these lines). This justifies what was said earlier—that an increase in lines alone could result in a worse picture! Experts generally say that the eye demands better definition horizontally than vertically.

For equal 625-line definition both ways the upper video frequency becomes approximately $600 \times 600 \times 25$ divided by 2, or 4·5Mc/s. Allowing for picture aspect (length to width) ratio and flyback time, the frequency becomes 5·5Mc/s or so (Western European stations are now transmitting a 5Mc/s video channel but 5·5Mc/s has been recommended by CCIR for u.h.f. transmissions). Dual-standard sets available seem to have a bandwidth of 4·25 to 5Mc/s when switched to 625 lines.

From the definition angle, therefore, the dual-standard switch has to change from about 2·7Mc/s video bandwidth and from 10,125c/s line frequency on 405 lines to about 5Mc/s video bandwidth and 15,625c/s line frequency on 625 lines.

In order to accommodate the higher video frequencies, therefore, the width of a 625-line channel is 8Mc/s, as compared with 5Mc/s in the 405-line case (see Fig. 1.1).

Modulation

The 625-line waveform itself differs in a number of fundamental ways from the original Blumlein 405-line waveform. A big difference is in the change from positive to negative vision modulation. In the 405-line waveform, sync. pulses are marked by a complete absence of modulation while in the 625-line waveform they appear as pulses of 100 per cent modulation (see Fig. 1.2).

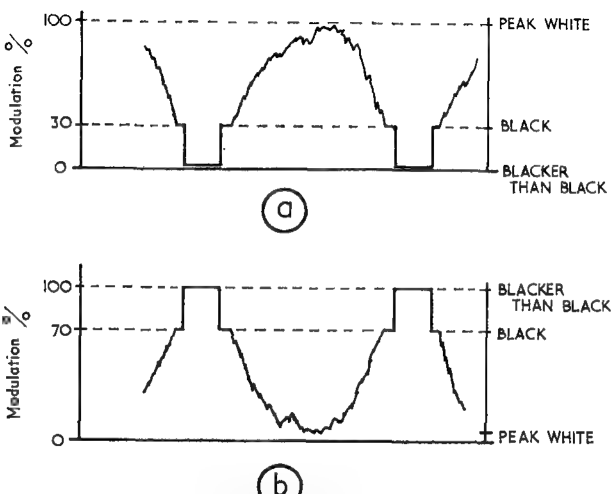


FIG. 1.2.(a) 405-LINE POSITIVE VISION MODULATION; AND (b) 625-LINE NEGATIVE VISION MODULATION. [NOTE THAT THE CARRIER NEVER FALLS TO ZERO]

On the sound side the modulation changes from a.m. on 405 lines to f.m. on 625 lines, and is spaced **3·5Mc/s below** the vision carrier in the former and **6Mc/s above** the vision carrier in the latter, as detailed in Fig. 1.1.

The "standard change" switch in a dual-standard receiver has between 15 and 20 "changeover" functions to look after, and is thus a multiple type of switch, usually of the slider variety mechanically coupled to the v.h.f. tuner, so that when turned to the "u.h.f." position the necessary 405 to 625 changeovers are accomplished.

The primary differences and the circuits which are switched on changing standard are given in Table 1.

TABLE 1

STANDARD	405 LINES	625 LINES	SWITCHING REQUIREMENTS
Frequency	V.H.F. (Bands I and III)	U.H.F. (Bands IV and V)	Change from v.h.f. to u.h.f. tuner
Video Bandwidth	2·7Mc/s	about 5Mc/s	Switching filters in the vision i.f. channel.
Sound Carrier Relative to Vision Carrier	−3·5Mc/s	+6Mc/s	Provided by tuner and vision i.f. switching.
Line Frequency	10,125c/s	15,625c/s	Line oscillator and output stage. Adding flywheel sync. on 625 lines.
Field Frequency	50	50	None.
Sense of Vision Modulation	positive	negative	Video amplifier and vision detector.
Sound Modulation	a.m.	f.m.	Sound detector. Also changing from 405-line sound i.f. to 6Mc/s intercarrier sound on 625 lines.

At this juncture it is as well to look at the four types of receiver which are currently in use. These are: (i) 405-line-only; (ii) convertible to dual-standard; (iii) dual-standard without u.h.f. tuner; and (iv) dual-standard with u.h.f. tuner fitted.

Those receivers in category (i) are not suitable for use on 625 lines. Those in (ii) were sold for immediate 405-line operation with some small or large provision for conversion to dual-standard working. A few models

in this category simply embody a two-position switch labelled "405-625". This switch may or may not be connected at the back. If it is connected then it probably just alters the line timebase. Other models are engineered to a degree where the 405-line panels can either be partnered with a set of 625-line panels or where the 405-line panels can easily be replaced by dual-standard panels. It is expected that service technicians and experimenters will be called upon to advise and modify such models over several years as the 625-line u.h.f. service spreads across the country starting from London, and then going on to the Isle of Wight, South Wales, the Midlands, South Yorkshire, Lancashire, Northern Ireland, Central Scotland, South-east England, Bristol, Northamptonshire, Suffolk, Norfolk, Nottinghamshire, North Yorkshire and Kincardineshire by 1967. The spread will continue after that date and there will be hosts of small booster stations to fill up the shadow areas with signal.

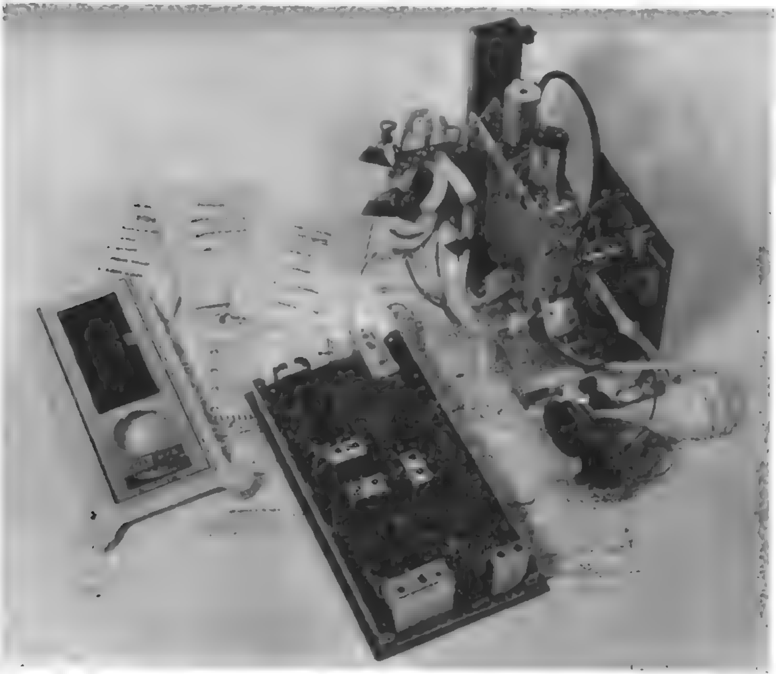


FIG. 1.3. ITEMS REQUIRED TO CONVERT CERTAIN COSSOR RECEIVERS FOR DUAL-STANDARD OPERATION

The items required for dual-standard conversion of the earlier Cossor range of receivers are shown in Fig. 1.3. A later chapter details conversion problems.

Models in category (iii) are already suitable for dual-standard operation but require the installation of a u.h.f. tuner. This is usually easily fitted by means of plugs and sockets. Category (iv) sets include the u.h.f. tuner and represent the true, practical dual-standard sets, it being just a

matter of connecting the v.h.f. and u.h.f. aerials to receive 405-line and 625-line pictures within the range of transmitters.

For the next few years sets in categories (i), (ii) and (iii) will continue to present a few problems. In areas embraced by the 625-line u.h.f. signal, sets mainly in category (iv) will be sold, while it is likely that the same models with or without the u.h.f. tuner (category (iii)) will be sold in other areas. The sale of 405-line-only models will speedily diminish.

2

THE U.H.F. TUNER AND FRONT-END COUPLING

DUAL-STANDARD receivers have two tuners: the ordinary switch or turret v.h.f. tuner; and a rather special tuner to select the u.h.f. channels. A u.h.f. tuner is shown on the right of Fig. 1.3. This differs mechanically from any v.h.f. tuner in that the tuning is continuously variable over the u.h.f. channels.

This is accomplished by a four-gang tuning capacitor assembly, the spindle of which is coupled to a slow-motion drive. The drive is in turn mechanically coupled to the tuning control knob and a channel indicator dial or scale. This is often reminiscent of the drive cord arrangement of radio sets.

The tuner uses two triode valves both operated in the earthed-grid mode, one as the u.h.f. amplifier and the other as the self-oscillating mixer (frequency-changer). The circuit of such a tuner is shown in Fig. 2.1. Here V_1 is the u.h.f. amplifier and V_2 the frequency-changer.

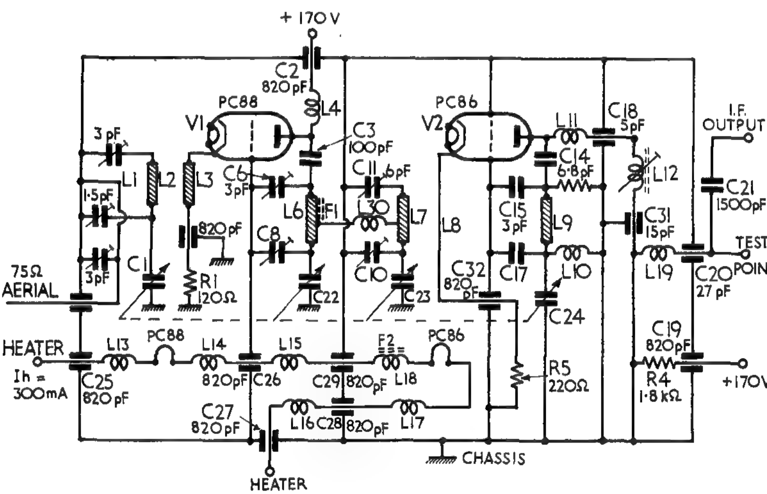


FIG. 2.1. CIRCUIT DIAGRAM OF THE MULLARD TYPE AT6360/02 U.H.F. TUNER. THIS, IN COMMON WITH OTHER MAKES OF U.H.F. TUNER, USES LECHER WIRES INSTEAD OF CONVENTIONAL COILS FOR THE TUNED CIRCUITS. THE FOUR MAIN LECHER WIRE CIRCUITS ARE TUNED BY SECTIONS OF A FOUR-GANGED CAPACITOR, AND THE TUNER IS BUILT INTO A THOROUGHLY SCREENED AND SEALED CONTAINER (SEE THE U.H.F. TUNER AT THE RIGHT OF FIG. 1.3)

Lecher Wires

An important feature of this kind of tuner is that it uses tuned lines (lecher wires) instead of conventional coils. At the frequency to which it is tuned, such a line produces a voltage/current distribution pattern exactly the same as that produced across a tuned half-wave dipole, as shown in Fig. 2.2.

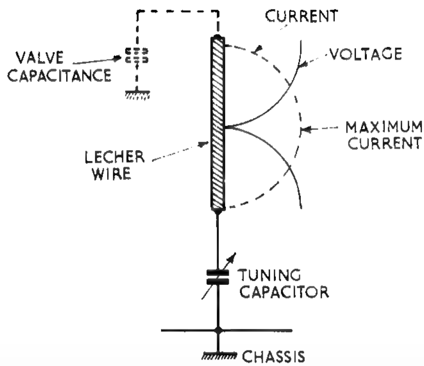


FIG. 2.2. THE VOLTAGE AND CURRENT DISTRIBUTION ACROSS A TUNED LECHER WIRE IS AKIN TO THAT ACROSS A TUNED HALF-WAVE DIPOLE AERIAL. THE WIRE IS LOADED AT ONE END BY THE VALVE OR CIRCUIT CAPACITANCE AND AT THE OTHER BY A SECTION OF THE TUNING GANG. ADJUSTING THE GANG EFFECTIVELY MOVES THE VOLTAGE/CURRENT PATTERN ALONG THE WIRE BUT WITHIN RANGE OF THE ASSOCIATED COUPLING

This technique is made possible since the length of a tuned line at u.h.f. can be accommodated within the small space of a tuner (*e.g.* a half-wave line at 960Mc/s is only about six inches long and the further physical length is shortened by capacitive loading at each end of the line). The lecher wires are loaded at one end by the tuning capacitor and at the other by the electrode capacitance of the valve. In effect, the tuning capacitor serves to vary the length of the line, and hence its tuned frequency. When the value of the capacitor is decreased, the tuned frequency increases, and both the voltage node and the current maximum (Fig. 2.2) move along the line towards the valve capacitance end, while the opposite movement takes place when the value of the tuning capacitance is increased. At mid-capacitance, the voltage and current are balanced.

Within their range of tuning, lecher wires create a surrounding field of signal energy which can be extracted either through an aperture or by way of a coupling loop. Signal energy can be introduced in a like manner.

For example, in Fig. 2.1 aerial signal is coupled *via* loop L_1 to lecher wire L_2. L_2 is then coupled to lecher wire L_3, which feeds the cathode of the u.h.f. amplifier valve. Capacitive and inductive values at u.h.f. are so small that just a protruding short length of wire is sufficient to couple the signal from one section to another.

The tuner is divided off into five channels or troughs with screening between, and each trough accommodates the appropriate lecher wire and section of the tuning gang and associated components. Coupling loops are positioned in the trough so that over the whole of the tuning range the voltage node and current maximum always remain within the coupling field. This ensures a relatively constant coupling in spite of the alteration in voltage and current distribution as the tuning is altered.

With all these factors in mind, we should now be able to understand the circuit in Fig. 2.1. Amplified u.h.f. signal at the anode of V_1 is developed across lecher wire L_6, and this in conjunction with L_7 forms

a bandpass-coupled circuit between the amplifier and the frequency-changer. This not only improves the selectivity and enhances the attenuation ratio with respect to signals falling within the image pass-band, but it also considerably reduces oscillator radiation from the aerial, a shortcoming which could otherwise prove embarrassing by pattern interference coming and going as neighbours tune over the u.h.f. channels.

Image Rejection

Early u.h.f. tuners had only a single-tuned coupling between the amplifier and the frequency-changer; but the resulting image attenuation was inadequate for the needs of the British TAC* 625-line system where a group of eleven consecutive u.h.f. channels is associated with each station area.

As a case in point, take a set tuned to Channel 23 in the London area. Here the vision carrier frequency is 487·25Mc/s. To give the preferred 625-line vision i.f. of 39·5Mc/s the local oscillator will be working at a frequency of 487·25+39·5Mc/s (as with v.h.f. the oscillator frequency is **above** the signal frequency), which is 526·75Mc/s.

Now, on Channel 33 (n+10) its vision carrier of 567·25Mc/s could produce a beat with the Channel 23 oscillator frequency, giving a difference-frequency (567·25−526·75Mc/s = 40·5Mc/s) which is only 1Mc/s removed from the vision i.f., but, of course, within the vision passband.

Here, then, is the reason for the good image rejection requirement. A rejection of about 53dB is used in Great Britain as compared with the Continental ratio of about 45dB.

To return to Fig. 2.1. The signal across lecher wire L_7 is coupled to loop L_8 which feeds the frequency-changer valve cathode. The oscillator operates in an "earthed-grid" Colpitts circuit, and tuning is performed by lecher wire L_9 which is similar to those used in the previous circuits. The 6·8pF (C_{14}) capacitor coupling the lecher wire to the anode of V_2 minimizes the detuning effect of the oscillator tuning capacitor on the i.f. output circuit. R.F. choke L_{10} is included for the same reason, effectively shorting C_{24} to i.f. signals.

To reduce radiation from the oscillator through the i.f. section, the i.f. circuits are connected to the frequency-changer anode through a low-pass filter, and link-coupling to the i.f. transformer is provided. A test point as well as the main i.f. signal output is available on most tuners. The i.f. tuned circuit is L_{12}.

It will be noticed that trimmers are connected to each end of lecher wires L_1, L_6 and L_7. These facilitate tracking of the ganged tuned circuits at the low and high frequency ends of the band. In effect, the capacitors near the gang end of the wires are adjusted for maximum high frequency end tracking while those at the opposite end of the wires are adjusted for optimum low frequency end tracking, just like the trimmers and padders in radio sets.

A word of warning at this juncture. U.H.F. tuners are not designed for general servicing, and the adjustments just described are, in fact,

* *TELEVISION ADVISORY COUNCIL* formed to advise on matters of planning, standards, etc.

carried out at the factory in conjunction with extremely accurate test-jigs, which cannot normally be reproduced in the service department.

Manufacturers strongly suggest that suspect and faulty u.h.f. tuners be returned to the factory for repair, for even removing and refitting the cover is likely to disturb severely the tracking and sensitivity.

To avoid chassis radiation the supply leads of the frequency-changer valve are passed through the various screening sections *via* feed-through capacitors. The cover is often lined with foam rubber and copper foil, thus ensuring a radiation-proof seal. On the Mullard tuner the chassis radiation specification is less than 50μV/metre, measured at a distance of 10 metres.

U.H.F. Valves

A popular valve for u.h.f. tuner service on the Continent is the **nuvistor**, which is a low-noise triode consisting of concentric cylinders sealed in an outer ceramic tube. In America disc-seal triodes of a similar nature and pencil valves are employed while in Great Britain major attention has been focused on the use of frame grid triodes.

These are typified by the Mullard PC86 (mixer/oscillator) and the PC88 (u.h.f. amplifier). They are mounted on B9A bases on which there is a sufficient number of pins to permit each electrode to have several connections.

From the grid aspect this is rather important, for instability can only be avoided in an earthed-grid stage by ensuring that the grid lead has the lowest possible value of inductance. This is achieved by the use of several grid connections.

Preferred I.F.

As already intimated, the preferred i.f. value for the British 625-line system is 39·5Mc/s for the vision carrier. Since the sound carrier is 6Mc/s above the vision carrier and the oscillator is above signal frequencies, the sound i.f. occurs at 33·5Mc/s. But on the 625-line system this is not fed through a sound i.f. channel, because both the vision and sound i.f. signals are fed together (at different levels) through the vision channel. At the vision detector the sound carrier beats with the vision carrier to form a 6Mc/s difference-frequency signal. This is frequency modulated with the sound signal and amplitude modulated with the vision signal. This so-called "intercarrier" signal can be used solely for the sound signal without vision interference due essentially to the different types of modulation used. More will be said later about the intercarrier signal.

Tuner Coupling

The i.f. signal from the output of the u.h.f. tuner is applied to the i.f. circuits in one of two ways. A popular arrangement is to feed the signal to the i.f. stages through the mixer section of the v.h.f. tuner, with the oscillator section muted. The v.h.f. mixer then serves as an extra stage of i.f. amplification at u.h.f., which is useful owing to the fact that the gain on the u.h.f. channels is usually less than the gain on the v.h.f. channels.

The second method is to apply the i.f. signals either from the v.h.f. tuner or u.h.f. tuner, as selected by a section of the standard change switch, direct to the input of the vision i.f. channel. Actually, on changing standards (as we shall see later), filter circuits are either switched into or removed from the vision i.f. channel to provide the requisite change in bandwidth. The tuner selecting switching is combined with the bandwidth switching operation.

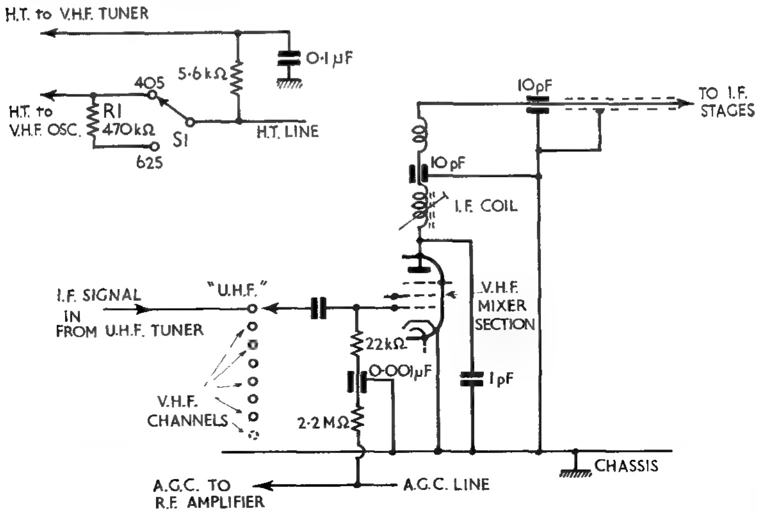


FIG. 2.3. A POPULAR ARRANGEMENT FOR COUPLING THE U.H.F. IF SIGNAL IS *VIA* THE MIXER STAGE OF THE V.H.F. TUNER. AN EXTRA POSITION MARKED "U.H.F." ON THE V.H.F. CHANNEL SELECTOR SWITCH MAKES THIS POSSIBLE. THE V.H.F. LOCAL OSCILLATOR IS MUTED ON THE U.H.F. SERVICE BY SWITCH *S1*. *R1* PASSES A SMALL OSCILLATOR VALVE ANODE CURRENT TO PREVENT CATHODE POISONING BUT INSUFFICIENT TO CAUSE OSCILLATION. THE EXTRA GAIN GIVEN BY THE MIXER ON U.H.F. COMPENSATES FOR THE LOWER GAIN AT U.H.F. AS COMPARED WITH V.H.F.

Fig. 2.3 shows the basic circuit when the u.h.f. tuner i.f. signals are applied to the v.h.f. mixer stage. When the v.h.f. channel selector is turned to the UHF position the u.h.f. i.f. signals are directed through a switch section to the control grid of the mixer pentode section of the frequency-changer valve and the normal mixer v.h.f. coil sections are removed.

Coupled to this switch position is the main standard change switch, and one section of this removes the h.t. voltage from the v.h.f. tuner, as shown by the second circuit in Fig. 2.3. When S_1 is in the "405" position, normal h.t. voltage is applied to the oscillator and the v.h.f. tuner works in the usual manner, while in the "625" position the 470-kΩ R_1 is added in series with the feed. This resistor cuts the voltage right down and prevents oscillation. The residual voltage is necessary to avoid "cathode poisoning" of the oscillator valve.

The i.f. output circuit is of the pi variety and h.t. voltage to the anode of the mixer valve is fed through the i.f. coupling coils in the i.f. channel proper.

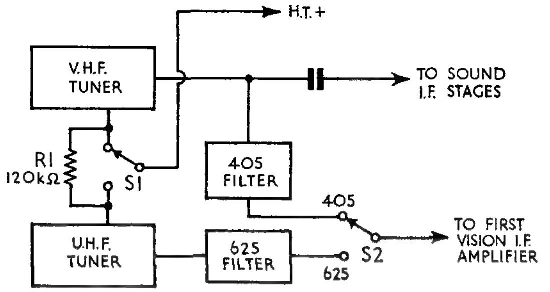


FIG. 2.4. ON SOME DUAL-STANDARD MODELS THE TUNERS ARE SWITCHED AS THIS DIAGRAM SHOWS. *R1* HERE SERVES THE SAME PURPOSE AS *R1* IN FIG. 2.3

It is usual to apply a.g.c. bias to the control grid of the mixer valve, as well as to the r.f. valve, when the v.h.f. tuner is used in this manner.

The arrangement for tuner switching is shown in Fig. 2.4. Here S_2 selects the appropriate tuner and filter coupling while S_1 changes over the h.t. feed. Resistor R_1 serves the same purpose as R_1 in Fig. 2.3. S_1 and S_2 are both sections of the dual-standard switching which are brought into operation when the v.h.f. tuner is switched to the U.H.F. position.

In Eire both 405 and 625 lines are accommodated on the v.h.f. channels so a u.h.f. tuner is not required. The v.h.f. tuner thus remains connected on both standards, even though the other circuits (detailed in Table 1, page 11) are changed over in the manner described in subsequent chapters.

U.H.F. tuners using transistors are common in the United States, but at the time of writing no British receiver adopts transistors in u.h.f. service. This will undoubtedly happen in the next year or so, however. Valve/transistor hybrid sets will also appear.

3

DUAL-STANDARD VISION SIGNAL STAGES

DUAL-STANDARD vision i.f. stages, taking in the vision detector and video amplifier stages, rate first in complexity so far as the standard switching is concerned. The i.f. stages themselves need almost to double in bandwidth on changing from 405 to 625 lines to embrace the top sidebands of the higher definition signals without attenuation. Moreover, the response must be wide enough to pass a little sound signal to provide the intercarrier signal from the beat of the vision and sound signals at the vision detector.

On the other hand, the 405 response must be tailored by rejectors at the sound i.f. (and at other frequencies) to prevent interference from the

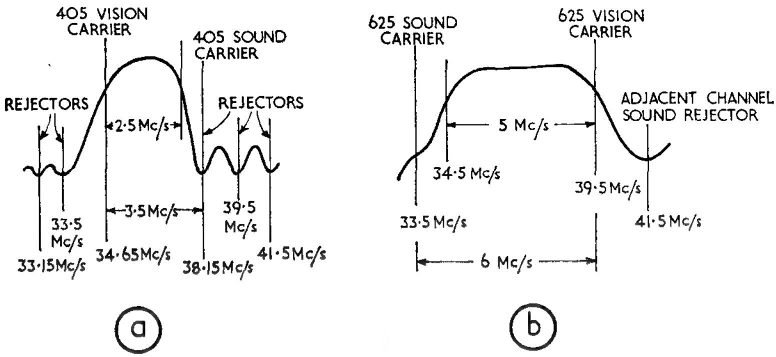


FIG. 3.1. REQUIRED 405 AND 625 RESPONSE CURVES AT (a) AND (b) RESPECTIVELY. DUAL-STANDARD I.F. STAGES ARE USUALLY DESIGNED FOR A 625 RESPONSE AND THEN TAILORED TO THE NARROWER 405 RESPONSE BY THE SWITCHING OF REJECTOR CIRCUITS

sound and adjacent signals. Then there is the reversal of the sound and vision carriers to contend with.

The required 405 and 625 response curves are depicted in Fig. 3.1 at (a) and (b) respectively. That at (a) is well known (apart, probably, from the troughs at the response limits due to the action of rejector circuits) while the curve at (b) is less familiar.

The i.f. stages are designed essentially to provide a passband of about 5Mc/s (curve (b)), and when the set is switched to 405 lines the tuner-to-i.f. coupling is modified and rejectors are switched to reduce the passband to something like 2·5Mc/s (curve (a)).

The circuit of the vision i.f. stages of the Ekco T398 dual-standard set is given in Fig. 3.2. Switches SW1J and SW1L are concerned with the response alteration, they being drawn in the "405" position.

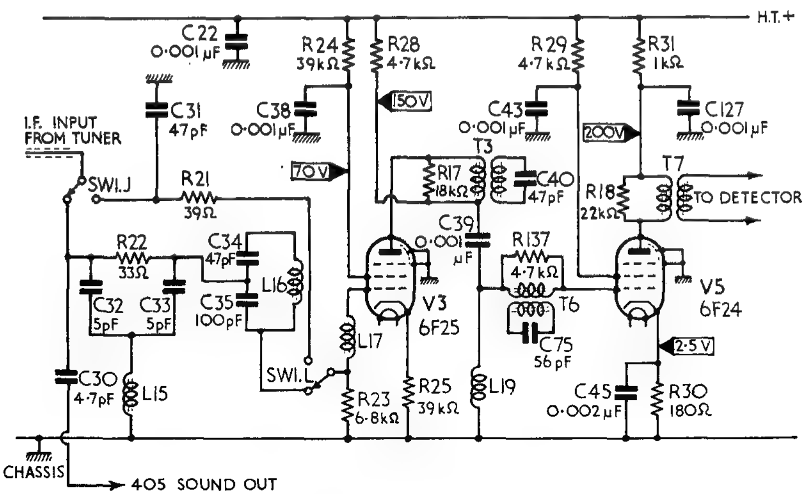


FIG. 3.2. DUAL-STANDARD I.F STAGE OF THE EKCO T398, DESCRIBED IN THE TEXT

On 405 lines the v.h.f. tuner i.f. signals are fed to the slider of SW1J, and the sound signal is passed out directly through the coupling capacitor C_{30}, and we shall later see what happens to this. The vision signal passes through filters comprising L_{15} and associated components and L_{16} and associated components. These circuits, in fact, constitute rejectors and they play a big part in the initial shaping of the 405-line vision response curve.

405 Response Shaping

L_{15} puts a trough in the response at the 405 sound i.f. of 38·15Mc/s and thus represents the ordinary sound rejector. L_{16} puts in a similar trough at 39·5Mc/s (the 625 vision i.f.), and the effects of both rejectors are revealed on the response curve in Fig. 3.1(a).

The troughs at 33·15Mc/s and 33·5Mc/s at the low frequency end of the 405 response curve are applied by "absorption" windings coupled to the i.f. output coil T_2 (see Fig. 3.3) of the v.h.f. tuner and coupling transformer T_3 (Fig. 3.2) respectively.

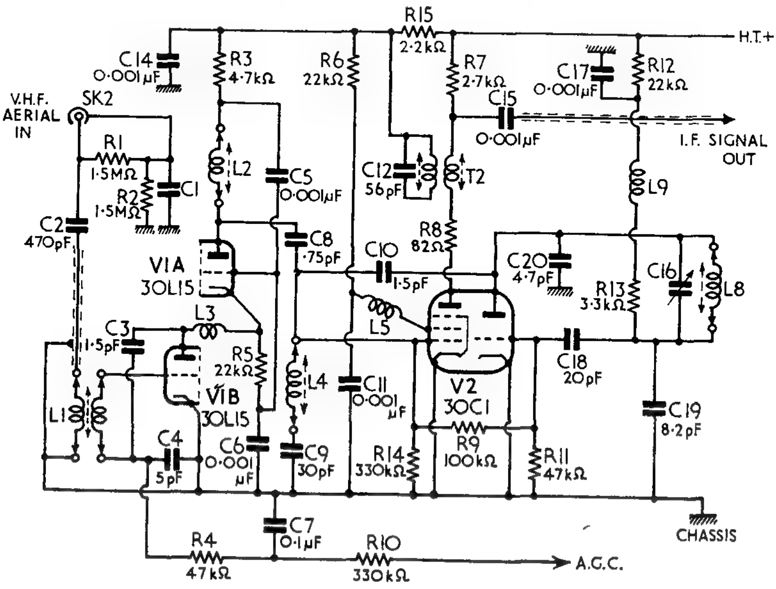


FIG. 3.3 THE V.H.F. TUNER I.F. COUPLING COIL IS OFTEN USED TO ADJUST THE RESPONSE ON 405 LINES. IN THIS V.H.F TUNER CIRCUIT THE WINDING COUPLED TO T2 ACTS AS AN ABSORPTION REJECTOR AT 33·15MC/S (SEE FIG. 3.1(a))

The trough at 41·5Mc/s is applied by the absorption winding coupled to T_6 (Fig. 3.2). These absorption windings are tuned to the rejection frequency and thus absorb signal energy at that frequency, thereby causing a dip or trough at the corresponding point on the response curve.

The rejection at 33·15Mc/s is for the prevention of breakthrough of the adjacent channel sound signal, separated by 1·5Mc/s from the vision carrier. That at 33·5Mc/s corresponds to the 625 sound carrier while

that at 41·5Mc/s corresponds to the adjacent channel sound signal on 625 lines. Other rejection frequencies are 38·15Mc/s 405-line sound carrier; and 39·65Mc/s 405-line adjacent channel vision carrier.

The general shaping of the body of the 405 response curve is accomplished by the tuned circuits L_{17}, T_3 and T_7 in the i.f. channel and T_2 in the tuner.

The rejectors at L_{15} and L_{16} are of the bridged-T type and thus provide a high degree of rejection at the critical frequency. Their adjustment is best undertaken by visual alignment procedures, using a wobbulator and oscilloscope, in conjunction with a marker generator.

When the receiver is switched to the "625" position switches SW1J and SW1L change-over. This action removes rejectors L_{15} and L_{16} and thus allows the high frequency end of the 405-line response to rise so as to give the passband shown in Fig. 3.1(b). The feed to the 405-line sound circuits is removed at the same time. The absorption rejectors retain the tailoring of the response limits in the former manner, but it is interesting to note that at the 625 sound frequency of 33·5Mc/s the response is not quite zero.

The shaping allows for a little of the sound signal to pass through the vision channel so that it can beat with the vision signal at the detector to produce the intercarrier sound signal, as previously mentioned.

The 625 sound signal is thus derived from the detector end of the vision channel, which is one of the reasons for the 405 sound circuit at the front of the circuit opening when the set is switched for 625-line working.

Points to note are the relative position of the 625 vision carrier when the response is opened out by the removal of the 405 rejectors and how the bandwidth is widened to 5Mc/s simply by the switching out of the rejectors. It is clear, therefore, that the circuits were designed to provide the 5Mc/s response curve as the first operation, with the second operation being to **narrow** the response to suit the 405 requirements with the minimum amount of switching.

Note also in Fig. 3.1 how neatly the reversal of relationship of the sound and vision carriers has been handled by the narrowing of the bandwidth due to the addition of filters on 405 lines.

Different models and makes adopt different methods of i.f. switching, but the net result is the same, as are also the frequencies involved.

Vision Detector

Fig. 3.4 shows the dual-standard vision detector circuit of the Ekco Model T398. The detector diode is MR_6 which is fed from the secondary of T_7 (Fig. 3.2). Now, on 405 lines switches SW1A and SW1C are as drawn in the circuit. The 405 detector load is then R_{38} and the i.f. filter (to prevent i.f. signal getting into the video stages) is L_{20}. The 405 vision signal is taken from across the load, *via* the coupling C_{54} and R_{39}, to the control grid of the video amplifier valve V_6, *via* the peaking coil L_{22}. This coil compensates for loss at high video frequencies.

The vision signal across the load R_{38} is positive-going. That is, it increases in a positive direction from black to white. The video amplifier valve is biased to accommodate this sort of signal by the cathode resistor

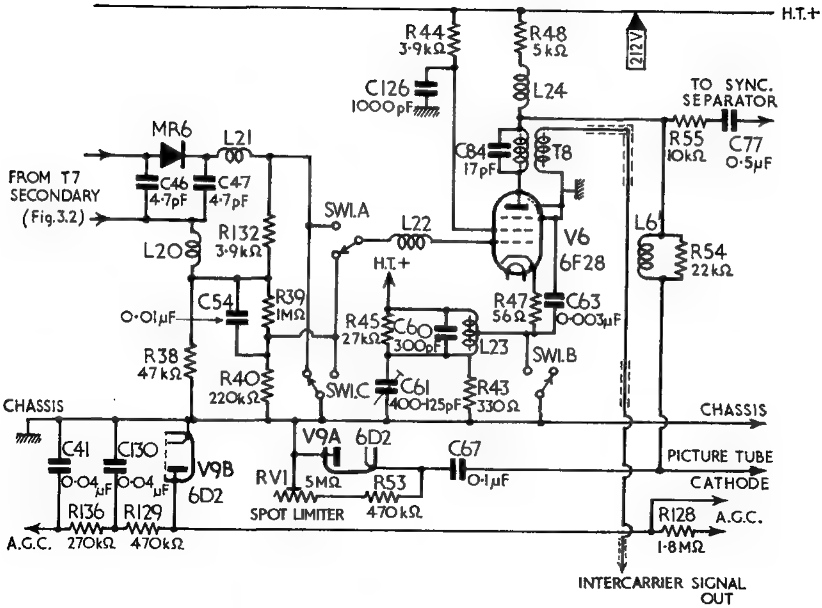


FIG. 3.4. DETECTOR, VIDEO AMPLIFIER, SPOT LIMITER AND A.G.C. FEED CIRCUITS OF EKCO DUAL-STANDARD RECEIVER, MODEL T398

network comprising R_{47}, R_{43} and R_{45}, and a negative-going vision signal is produced across the anode load R_{48}. This is fed out through the peaking coil L_6 and the parallel damping resistor (to prevent "ringing") to the cathode of the picture tube, *via* a d.c. component reduction circuit made up of a 100-kΩ resistor in parallel with a 0·1-μF capacitor. The action of this circuit is to reduce the d.c. signal component to avoid aircraft flutter and certain other a.g.c. effects.

Tuned circuit L_{23} and C_{60} in V_6 cathode circuit is to reject 3·5Mc/s line break-up due to the beat of the 405 sound and vision carriers. The tuning is adjusted to eliminate the "dot pattern" on the scanning lines.

Video compensation is given by the trimmer C_{61}, which is adjusted for optimum definition, it adjusting the negative feedback in the video amplifier, and L_{24} in the anode circuit of V_6. This is really a video signal peaking coil since it tends to neutralize the capacitive loading which, of course, would otherwise attenuate the higher video frequency signals and impair the picture definition. Such video compensation is used not only in dual-standard models, but is also found in the majority of 405-line-only and 625-line-only models.

Action of Video Amplifier

Fig. 3.5 shows the r.f. waveform of the radiated 625 and 405 signals at (a) and (b) respectively. At (a) maximum transmitter power occurs on the sync. pulses while at (b) it occurs at peak white picture (see also Fig. 1.2, page 10). In spite of these differences, the cathode of the picture

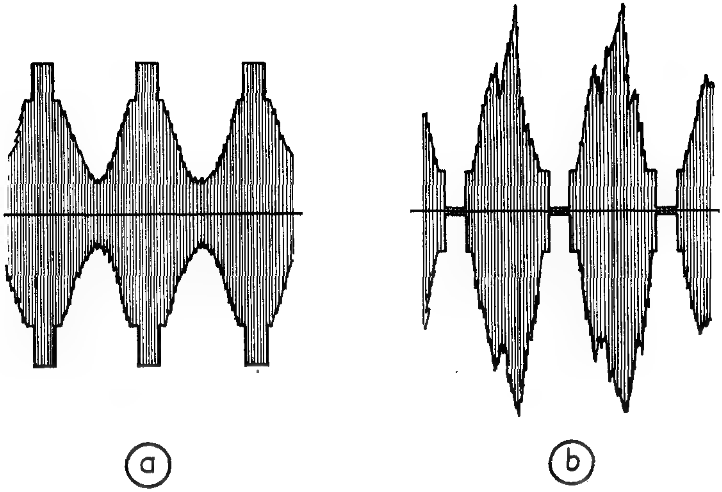


FIG. 3.5. RF WAVEFORMS OF RADIATED 625 AND 405 SIGNALS AT (a) AND (b) RESPECTIVELY

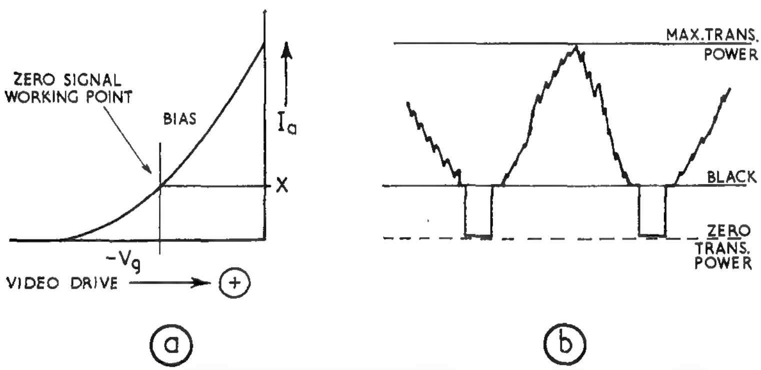


FIG. 3.6. DEMODULATED 405 VISION SIGNAL (b). AT (a) IS SHOWN THE REQUIREMENTS OF THE VIDEO AMPLIFIER VALVE TO HANDLE SUCH A SIGNAL. NOTE THE POSITIVE-GOING VIDEO DRIVE CAUSING A RISE IN ANODE CURRENT FROM BLACK TO WHITE SIGNAL

tube must "see" a negative-going picture signal (this is the same as the grid going positive with respect to cathode). This means that the control grid of a single-stage video amplifier must "see" a positive-going picture signal, since the valve reverses the phase of the signal.

From the aspect of 405 positive vision modulation, the detector will produce across its load a demodulated signal after the style of that depicted at (b) in Fig. 3.6. We have seen that the video amplifier is biased to handle such a signal and at (a) is shown how the working point for zero signal is set by the bias. As the signal rises positively, so the video amplifier valve anode current rises and the voltage across the anode load rises negatively. That is, the voltage at the anode effectively falls as

the anode current rises with increase in white signal. Thus we have a negative-going signal.

To see what happens on 625 lines we must look back again to the circuit in Fig. 3.4. On the "625" position switches SW1C, SW1A and SW1B change over. SW1C removes the chassis connection from the "cathode" side of the detector diode MR_6 and, from the signal aspect (taking C_{54} into consideration), "earths" the bottom end of R_{132}, which then becomes the 625 detector load resistor.

SW1A changes so that the signal across the new load is fed to the video amplifier valve control grid. These two actions deal with the reversal of the phase of the vision detector, this being necessary to cater for the change in the polarity of the vision signal.

Different models and makes incorporate slightly different methods of accomplishing this phasing change, but the net result is always the same.

Switch section SW1B is concerned with the biasing of the video amplifier valve, for when the polarity is changed, the video drive conditions to the amplifier may also change, as we shall see later.

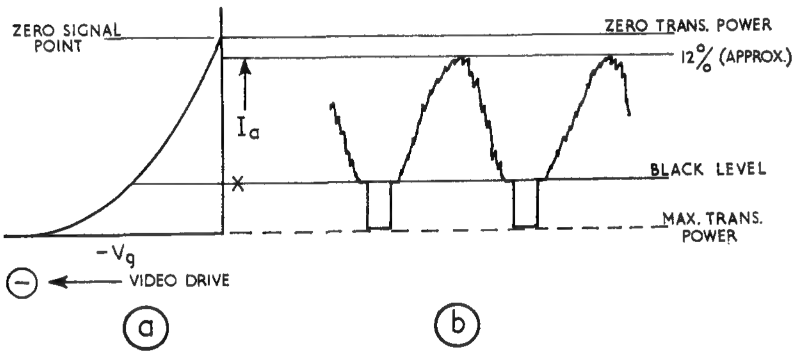


FIG. 3.7. DEMODULATED 625 VISION SIGNAL (b). AT (a) IS SHOWN THE REQUIREMENTS OF THE VIDEO AMPLIFIER VALVE TO HANDLE SUCH A SIGNAL. AT ZERO SIGNAL ACROSS THE DETECTOR LOAD THE VALVE ANODE CURRENT IS HIGH WHILE AT BLACK LEVEL IT IS BIASED TO CONDITIONS COMPARABLE TO THOSE IN THE 405-LINE CASE. NOTE THAT THE ANODE CURRENT IS CAUSED TO RISE FROM BLACK LEVEL TO WHITE SIGNAL BECAUSE OF THE VIDEO DRIVE GOING *LESS NEGATIVE*. THIS IS EQUIVALENT TO THE 405-LINE CASE WHERE THE VIDEO DRIVE GOES *MORE POSITIVE*. THE PICTURE TUBE THUS "SEES" A NEGATIVE-GOING PICTURE SIGNAL IN EACH CASE

The switch shorts out the biasing resistor R_{43} and the 3·5Mc/s "dot pattern" rejector which is redundant on the 625 standard, the spacing between the sound and vision carriers having changed from 3·5Mc/s to 6Mc/s, the latter which we need to preserve and not reject for it is this that gives the intercarrier sound signal.

There are various ways of showing the change in video standard. One is shown in Fig. 3·7, which should be compared with that in Fig. 3.6. On both standards there is zero voltage across the detector load under conditions of zero signal. In Fig. 3.6 we have seen that the video amplifier valve is biased to low current for zero signal. Fig. 3.7, however, shows that for zero signal there is high current in the amplifier (remember we have shorted out most of the cathode bias resistor).

At signal black level the anode current **rises** to point *X* in Fig. 3.6(a) and **falls** to point *X* in Fig. 3.7(a). The conditions at black signal are thus comparable on both systems. Note that the video drive on 405 lines has risen **positively** from zero signal to black (Fig. 3.6(b)), while on 625 lines it has risen negatively (Fig. 3.7(b)).

From black level towards white the 405 video drive goes **more positive**, thereby increasing the video amplifier valve anode current. On 625 lines the drive goes **less negative**. This, of course, is the same as if it were said to be going more positive, which means that the video valve anode current also increases from signal black towards white on the 625 standard.

Signal-wise, the conditions are thus similar on both standards, in spite of the change in modulation polarity, and the tube cathode "sees" a negative-going signal in each case.

D.C. Coupling

From Fig. 3.4 it will be seen that the d.c. component of the video signal is retained right from the vision detector load to the cathode of the picture tube. This follows conventional 405-line practice, though the d.c. level is often purposely attenuated by the use of a capacitor in parallel with a resistor in the feed circuit from the video amplifier valve anode to the tube cathode. The capacitor passes video signal above d.c. in a rising response while the resistor introduces the requisite degree of d.c. coupling.

D.C. attenuation tends to reduce the effects of aircraft flutter and other shortcomings which arise from slight variations in the d.c. conditions.

Moreover, the application of mean level vision a.g.c., from the grid circuit of the sync. separator valve, itself tends to cancel full d.c. coupling which may be used from detector to tube cathode. It is thus the practice to attenuate the d.c.

On some dual-standard models, however, the d.c. coupling is completely removed on the "625" position. How this is accomplished is shown in Fig. 3.8. On 405 lines S_1 is closed to short circuit C_1. The normal d.c. coupling is then adopted between the detector and the video amplifier valve grid, and the amplifier is biased by the cathode resistor R_1.

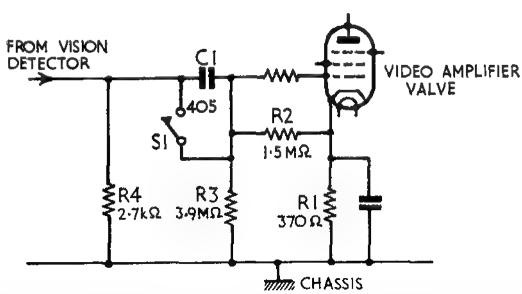


FIG. 3.8. IN SOME MODELS THE NORMAL 405 D.C. COUPLING BETWEEN THE VISION DETECTOR AND THE VIDEO AMPLIFIER MAY BE DESTROYED BY THE SWITCHING-IN OF A COUPLING CAPACITOR, SUCH AS *C1*. THE VIDEO AMPLIFIER BIASING IS AT THE SAME TIME ALTERED SLIGHTLY, BY *R4* BEING REMOVED FROM THE JUNCTION OF *R2 R3* BY *S1* OPENING. THIS TECHNIQUE AVOIDS HIGH ANODE CURRENT IN THE VIDEO AMPLIFIER VALVE AT ZERO SIGNAL, SINCE THE VIDEO AMPLIFIER MAY BE BIASED TOWARDS CLASS-A CONDITIONS ON BOTH STANDARDS

On the "625" position S_1 opens and interposes C_1 in the signal coupling. The d.c. coupling is then eliminated. In addition, a small modification to the amplifier bias occurs due to the effect of the cathode potential-divider network R_2,R_3. This network reduces the bias on 625 lines, but owing to the relatively low value of R_4 shunting the network when S_1 is closed it has little effect on 405 lines.

By changing to a.c. coupling on 625 lines the video amplifier biasing, in general, needs only a small change to accommodate both standards and it is not necessary to switch the cathode circuit. Moreover, since the valve operates towards Class-A in both cases, the anode current at zero 625 signal does not rise to a high level. Both arrangements are found in practice.

Intercarrier Signal

The 6Mc/s beat between the sound and vision 625 carriers at the vision detector is developed in terms of a signal across a 6Mc/s tuned circuit either directly after the vision detector or in the anode circuit of the video amplifier valve. Fig. 3.9 shows how the signal is extracted directly following

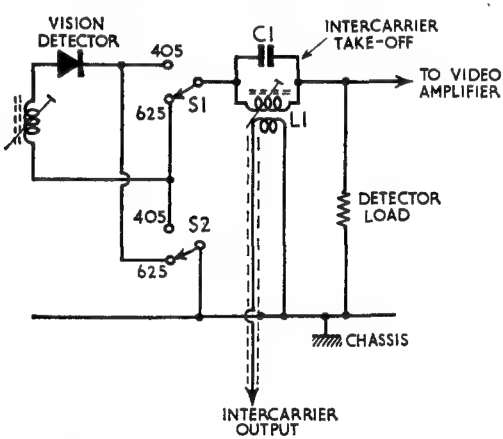


FIG. 3.9. SHOWING HOW THE INTERCARRIER SIGNAL IS SOMETIMES EXTRACTED AFTER THE VISION DETECTOR. *C1/L1* CIRCUIT IS RESONATED TO 6MC/S AND A LOW IMPEDANCE COUPLING WINDING CONVEYS THE SIGNAL TO THE SOUND CHANNEL. ON 405 LINES A CHANGE-OVER SWITCH MAY SHORT THE INTERCARRIER FEED THROUGH A CAPACITOR (*E.G., S1* AND *C69* OF FIG. 3.10). THE TUNED CIRCUIT ALSO SERVES AS A REJECTOR AND PREVENTS THE INTERCARRIER SIGNAL FROM ENTERING THE VIDEO AMPLIFIER. THE CIRCUIT DOES NOT SHOW THE FILTER CHOKES, BUT IT DOES ILLUSTRATE AN ADDITIONAL METHOD OF CHANGING THE PHASE OF THE VISION DETECTOR IN CONJUNCTION WITH A COMMON LOAD RESISTOR AND SWITCHES *S1* AND *S2*. WHEN THE SIGNAL IS TAKEN OUT AFTER THE VIDEO AMPLIFIER, THE TUNED COUPLING CIRCUIT IS INTERPOSED IN THE VALVE ANODE CIRCUIT (*E.G., T8* AND *C84* IN FIG. 3.4)

the vision detector, and this circuit, incidentally, shows another method which is sometimes used to change the phase of the detector, using a common load resistor.

Filter L_1,C_1 is tuned to 6Mc/s and a low impedance coupling coil is used to extract the intercarrier signal. The tuned circuit offers a very low series impedance to the video signal, so that this is unhindered in passing from the detector to the amplifier. However, the circuit offers a

high impedance at 6Mc/s and thus prevents the intercarrier signal from reaching the video amplifier.

When the intercarrier signal is taken from the anode circuit of the video amplifier valve, it is allowed to pass into the grid circuit along with the video signal. The intercarrier signal is then developed across a similar tuned circuit in the anode. In Fig. 3.4 this circuit comprises C_{84} across the primary of T_8, with the secondary forming a low impedance coupling output. The primary of T_8 also acts as a 6Mc/s rejector in the video amplifier circuit, thereby preventing the intercarrier signal from reaching the tube cathode.

The advantage of passing the intercarrier signal through the video amplifier is that extra lift is given to the signal before it is fed to the sound channel, meaning that the sound channel need have less 6Mc/s gain than when the signal is extracted from the detector output.

A disadvantage is intermodulation resulting from overloading in the video amplifier.

Some colour television sets using the intercarrier sound technique incorporate a separate detector at the output of the vision i.f. channel solely for producing the intercarrier sound signal. In that way the 6Mc/s beat can be directed away from the vision detector proper. This scheme makes it easier to suppress the intercarrier signal from the video circuits where, in colour sets in particular, secondary beating may occur with the colour subcarrier and cause a dot pattern effect which appears to be "crawling" over the picture.

It will be understood, of course, that the intercarrier signal has signal components which are both amplitude and frequency modulated. The sound signal is frequency modulated, and as the sound detector should be sensitive only to an f.m. signal, the amplitude modulation of the vision signal component should be completely suppressed.

Intercarrier Buzz

The main a.m. component is at the 50c/s field frequency, and if the a.m. rejection is not as good as it should be a "buzz" appears on the sound. On the 625 standard this is called **intercarrier buzz.**

It will be recalled that a similar effect can occur on the 405 standard due to vision signal breakthrough into the sound channel, called **vision-on-sound**. The main causes of the trouble in the 405-line case are misalignment of the sound i.f. channel and intermodulation of the sound and vision signals due to overloading and resulting non-linearity in a stage common to both sound and vision signals.

Apart from impaired a.m. limiting in the 625 f.m. detector stage, incorrect alignment of the vision i.f. stages can also aggravate intercarrier buzz. This can happen if the alignment is such that insufficient level of sound signal arrives at the vision detector. A rejector or tuned circuit of some kind is included in the vision i.f. channel on 625 lines to hold the sound carrier (33·5Mc/s) at the most desirable level (see Fig. 3.1(b)) so far as the production of a noise-free and distortion-free intercarrier signal is concerned.

The sound signal is typically about 30dB below the level of the vision signal at the vision detector. If the sound signal rises much above this

level, sound distortion can prove troublesome. The correct level is established during the process of alignment as recommended by the manufacturer. Here, then, is an extra factor illustrating the importance of correct alignment.

In addition to the a.m. rejection given by the f.m. detector, most models include some form of a.m. limiting in the sound channel on the 625 standard.

One of the advantages of the intercarrier sound technique is that sound fade which would otherwise occur due to drift of the u.h.f. local oscillator is avoided. The intercarrier frequency is held accurately at the transmitter by the difference in frequency between the sound and vision carriers. Alteration in frequency of the local oscillator does not alter this difference-frequency, as it alters the sound i.f. in the 405-line case.

Thus, provided the sound and vision carriers fall within the 625 i.f. passband an intercarrier signal will be produced. However, distortion is likely to result if the carriers are misplaced on the response curve due to maladjustment of the fine tuning control (u.h.f. channel selector control knob) or misalignment.

The Sound Channel

The sound i.f. channel of a dual-standard set is designed to be responsive to the standard a.m. sound i.f. (38·15Mc/s) and also to the 6Mc/s intercarrier sound signal with very little or no switching at all.

The change in response usually requires no switching owing to the wide difference between the frequencies of the two signals concerned. Series-connected i.f./intercarrier transformers are employed, as shown overleaf in Fig. 3.10.

Here the tuner signal is applied to L_3, *via* C_7, while the vision i.f. signal is passed out through L_1 to the vision i.f. channel. L_3 has two tuned circuits connected in series. Circuit (A) with C_8 is tuned to the 405 sound (38·15Mc/s) while circuit (B) with C_9 is tuned to the intercarrier sound (6Mc/s).

On 405 lines the intercarrier take-off is short circuited through S_1 and C_{69} (extreme right of Fig. 3.10).

The input 38·15Mc/s 405 sound signal is thus developed across L_3A, and after amplification by V_1 is developed across the coupling transformer T_1. The signal is further amplified by V_2 and finally developed across T_4, the transformer which feeds the 405 a.m. sound detector D_1.

The series-connected 6Mc/s tuned circuits do not detract from the efficiency, since at 38·15Mc/s their parallel capacitances represent very low impedances.

On 625 lines, the 6Mc/s intercarrier signal is effectively applied across L_3B (Fig. 3.10). This signal is amplified, as in the 405-line case, by V_1 and developed across T_2 and then by V_2 and developed across T_5, the ratio detector transformer for feeding the associated diodes D_2 and D_3.

The series-connected 38·15Mc/s tuned circuits do not detract from the efficiency, since at 6Mc/s their small inductance values represent low impedances.

In this way, then, a sound section which is responsive to both 38·15Mc/s and 6Mc/s signals is available without switching. It will be

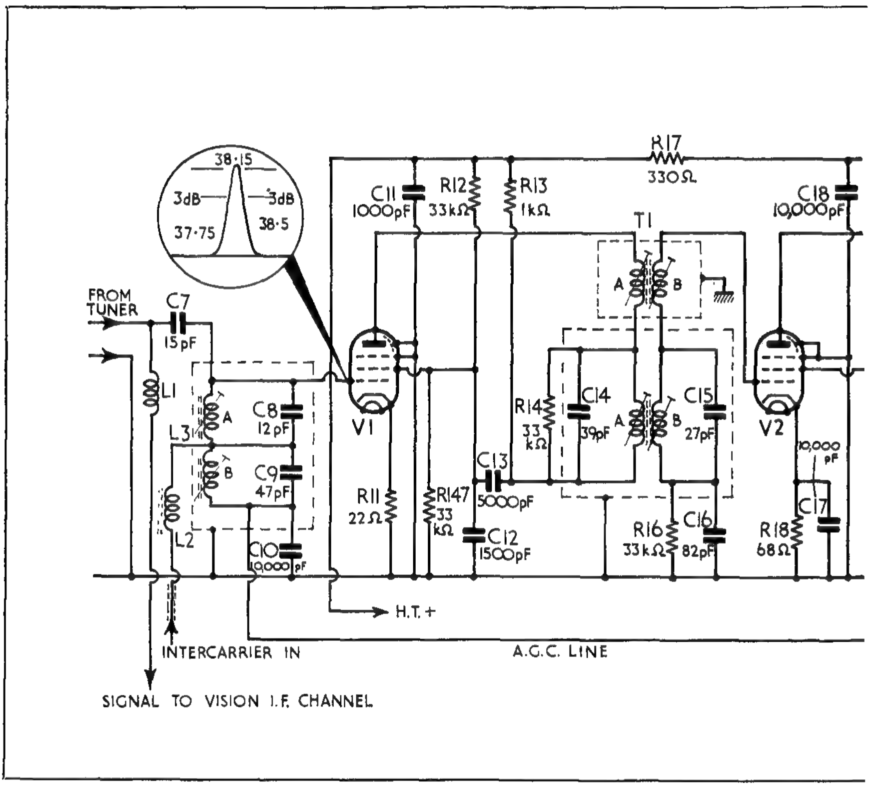
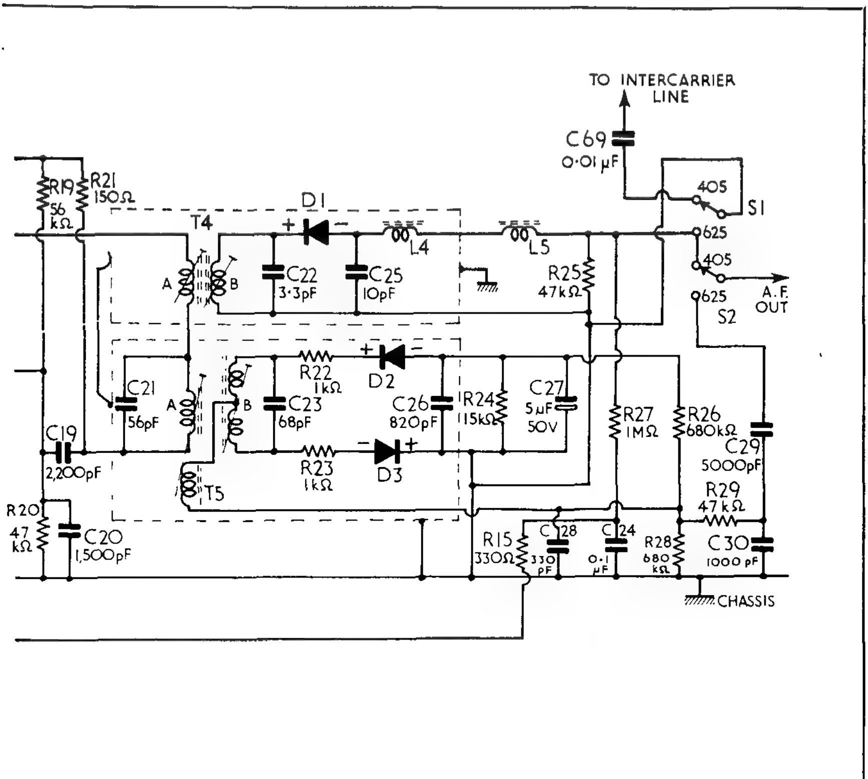


FIG. 3.10. THE SOUND CHANNEL OF A DUAL-STANDARD RECEIVER,



SHOWING THE A.M. AND F.M. DETECTORS AND SWITCHING

recalled that the same technique is adopted in a.m./f.m. radio receivers. The overall 38·15Mc/s response of the channel is shown in Fig. 3.10. This would be obtained by feeding a wobbulated signal in at V_1 control grid and obtaining the display from the a.m. detector load.

A.M. Detector and A.G.C.

The a.m. 405 sound detector is conventional. R_{25} is the load, with C_{25}, L_4 and L_5 as filter components to block the 38·15Mc/s signal. Audio across the load is fed to the a.f. section *via* S_2. S_1 shorts out the inter-carrier feed on 405 lines.

The negative potential at the "live" side of the load resistor R_{25} is used as an a.g.c. bias. It is filtered by R_{27} and C_{24} and fed to the control grid of V_1 through R_{15} and L_3A/B. As the negative potential increases with increase in signal strength, the gain of V_1 is automatically pulled down with increase in signal level, owing to the negative bias reducing the effective mutual conductance of the valve.

F.M. Detector and Limiting

The ratio detector in Fig. 3.10 is also fairly conventional. T_5 is the ratio detector transformer, with the normal third (tertiary) winding for phase reference. The ratio detector diodes are D_2 and D_3 while the d.c. load and stabilizing capacitor are R_{24} and C_{27} respectively.

The f.m.-derived audio is developed across C_{28} and R_{28} connected to one end of the tertiary winding. Normal pre-emphasis is given at the station, so receiver de-emphasis is necessary and is given by R_{29} and C_{30}. The corrected signal is then fed through C_{29} to the audio section, *via* S_2.

Note that S_1 shorts out the a.m. detector load and a.g.c. in the "625" position. This means that on 625 lines V_1 is always at maximum gain. The intercarrier signal is generally at a fairly high level so that V_1 is almost fully loaded. V_2 is, in fact, pushed hard due to the amplified signal and it tends to limit. This action is assisted by the grid current components R_{16} and C_{16}.

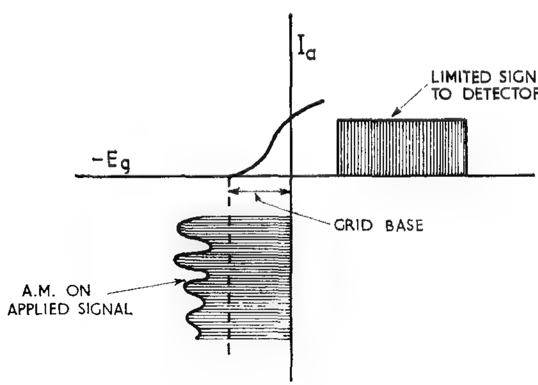


FIG. 3.11. SHOWING HOW A SHORT GRID BASE OF A PENTODE I.F. AMPLIFIER CAN LIMIT THE AMPLITUDE OF THE 625 SOUND CARRIER

V_2 is normally operated with a short grid base due to the low and stabilized screen voltage, given by the potential-divider R_{19}-R_{20}. When the signal amplitude exceeds the grid base of the valve, grid current flows in R_{16} and C_{16} charges so that the grid is biased negatively. If the amplitude of the signal tends to increase, the negative grid bias will also increase, and the effect will be to hold the output signal constant.

This amplitude limiting action shaves off interference pulses and amplitude fluctuations of the f.m. signal before it is applied to the detector. The f.m. detector, of course, is not concerned with amplitude modulation and any such disturbance of the signal is liable to provoke unnecessary noise and interference (such as intercarrier buzz, previously referred to). It is thus permissible (and desirable) to limit the amplitude of the signal on the 625-line standard. The action is shown diagrammatically in Fig. 3.11.

On the 405-line standard, of course, such limiting cannot be employed, since the audio is carried by virtue of amplitude variations of the sound carrier.

Ratio Detector Limiting

The ratio detector itself also possesses an inherent amplitude limiting attribute. Referring to Fig. 3.10, diodes D_2 and D_3, being series-connected, rectify the signal and produce a d.c. potential across R_{24}. This charges C_{27}, and the value of this *CR* time-constant is typically 0·2 seconds. Thus, rapid variation of the amplitude of the input signal ceases before it can influence the charge on C_{27}.

An input signal of constant amplitude gives a settled charge on C_{27}, but when the amplitude of the signal increases suddenly the voltage across R_{24} tends to rise and as a consequence current flows through R_{24} into C_{27} to increase the latter's charge. This current flow "looks" to the diode circuit as a sudden decrease in the value of the load and the effective damping across the tuned circuit rises. This tends to offset the amplitude

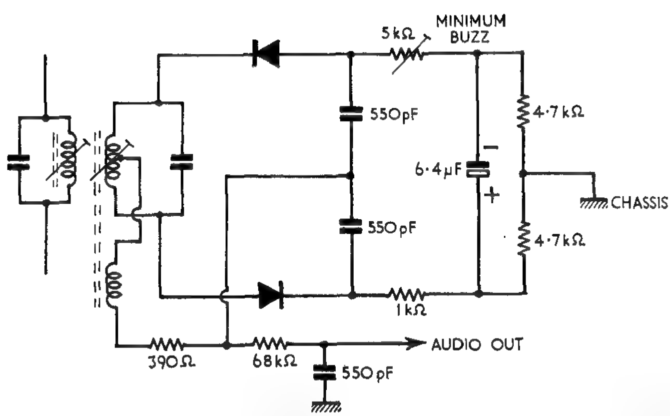


FIG. 3.12. A BALANCED RATIO DETECTOR CIRCUIT. THE 5-KΩ "MINIMUM BUZZ" PRESET IS USED TO BALANCE THE CIRCUIT AND ALLEVIATE INTERCARRIER BUZZ

increase of the input signal and consequently has a stabilizing effect on the output voltage.

A sudden decrease in the amplitude of the applied signal has the converse effect, *i.e.* the tuned circuit loading is lightened by current flowing out of the capacitor.

How well a ratio detector limits is also influenced by the balance of the circuit. Unbalance due to misalignment or due to unmatched diodes or associated components allows the detector to respond to amplitude modulation.

The f.m. detector in some dual-standard models features a preset resistor which can be used to effect optimum detector balance. It is usually adjusted for minimum intercarrier buzz (Fig. 3.12).

Locked Oscillator f.m. Detector

Another form of f.m. detector currently employed in dual-standard models uses a Mullard heptode EH90 valve. The circuit is shown in Fig. 3.13.

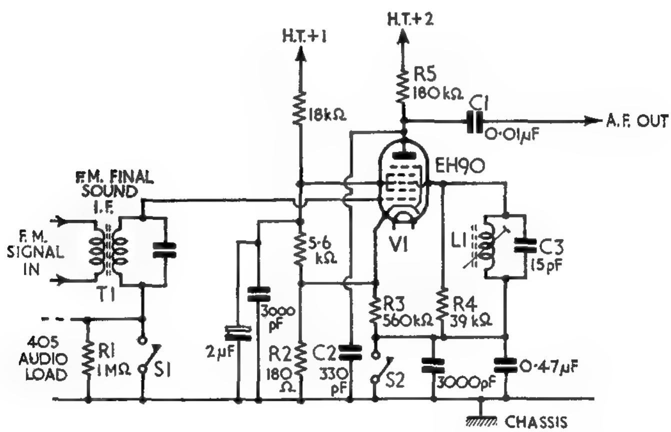


FIG. 3.13. CIRCUIT OF F.M. DETECTOR/A.F. AMPLIFIER USING THE EH90 HEPTODE VALVE. A FULL DESCRIPTION OF THE OPERATION OF THE CIRCUIT IS GIVEN IN THE TEXT

On 405 lines the valve V_1 acts as an ordinary a.f. amplifier. Signal across the audio load resistor R_1 is fed through the secondary of T_2 to grid 1 of V_1, since S_1 is open.

S_2 is closed, which means that grid 3 leak resistor R_4 is connected to chassis and the full bias as developed across R_2 is applied to grid 3. The shunting effect of R_3 is negligible since its value is high compared with that of R_2. Grids 2 and 4 are rather like screen grids and are connected to HT+1, *via* a potential-divider network. Audio signal is thus developed across the anode load resistor R_5 and is fed through C_1 to the audio stages.

Grids 1 and 3 of V_1 have a very short grid base. That is, anode current reaches saturation with a small positive signal swing on either grid while it goes into cut-off with a small negative signal swing on either grid. The

difference in grid voltage between the conditions of saturation and cut-off is a measure of the grid base of the valve (also see Fig. 3.11).

Grids 1 and 3 can really be considered as control grids since the anode current is readily influenced by the bias on both or either. Anode current flows only when both grids are suitably biased away from cut-off. If any one grid is biased away from cut-off while the other grid is biased for cut-off anode current will not flow.

In Fig. 3.13 the 6Mc/s f.m. carrier is applied from T_1 to grid 1. Now, let it be supposed that grid 3 is biased away from anode current cut-off while grid 1 is biased towards the centre of its small range of operational swing. Thus, when the f.m. carrier swings positive and negative over a complete cycle the valve is pushed alternately into saturation and anode current cut-off.

This results in both the positive and negative tips of the signal being sliced off, giving approximately squarewave signal pulses across the load resistor R_5. The amplitude of these pulses would be held constant even though the input signal may rise in amplitude, due to the short grid base of grid 1. The valve thus acts as a good amplitude limiter.

These signal pulses cause C_2 (connected between the anode and chassis) to charge, but since the amplitude and duration of the pulses are constant, the voltage of the charge across C_2 also remains constant.

Before the valve serves as an f.m. detector other things must happen. Let us now suppose that grid 1 is biased away from anode current cut-off while grid 3 has applied to it an oscillatory voltage. The effect in the anode circuit would be exactly the same as described for grid 1.

In practice, grid 3 is given an oscillatory voltage, since L_1 and C_3 form an oscillator tuned circuit and the valve is caused to oscillate at 6Mc/s due to coupling between the cathode (S_2 opens on 625 lines) and grid 3. If we still consider that grid 1 is biased away from anode current cut-off and is without a signal of its own, then the pulses across the anode load R_5 due to the oscillatory voltage at grid 3 again charge C_2 to a constant value.

We must now consider the action with the 6Mc/s, unmodulated intercarrier signal applied to grid 1 at the same time as the oscillatory voltage is applied to grid 3. Due to electron coupling within the valve, the oscillator locks to the 6Mc/s intercarrier frequency over a small range of frequencies, called the **deviation range** or **pull-in range.** Outside this range the oscillator runs freely and beats occur between the input and oscillator signals, as would be expected. The tuned circuit L_1-C_3 must, therefore, be tuned closely to 6Mc/s.

When the oscillator is in this way "synchronized" to the intercarrier signal grid 1 swings positive during the period that grid 3 swings positive, but owing to the space charge (or electron) coupling within the valve the voltage at grid 3 is 90° out of phase with the voltage at grid 1.

The effect in terms of pulses of anode current is that each pulse is half the width of a pulse which would result if only one grid were in receipt of a signal while the other grid were biased away from anode current cut-off, as we have already considered. This is because the two pulses are displaced by 90° from each other (a quarter of a cycle), bearing in mind that each pulse is representative of a half cycle of signal under single signal conditions.

The two pulses are superimposed over half their width which produces the pulse of half width. If the signal voltages at both grids were exactly in phase, a full-width pulse would occur while if they were exactly anti-phase (180°), there would be no pulse at all. Actually, under these extreme conditions the oscillator would fall out of lock from the intercarrier signal.

During the normal f.m. deviation when the intercarrier signal is modulated, the oscillator is held within the deviation range and it remains locked to the signal. During this time, however, the phase between the signal voltages at grids 1 and 2 varies either side of the nominal 90° value for an unmodulated signal.

This means that the effective width of the pulses across the anode load resistor varies in accordance with the modulation depth (or deviation). The variation is either side of the quarter-cycle width as the modulation causes the signal on grid 1 to increase and decrease in frequency up to a maximum deviation of $\pm$75kc/s.

Although the amplitude of the pulses remains constant due to the limiting action of the valve, the changing pulse width results in a corresponding change in charge across the anode capacitor C_2. This integrates (adds) the pulses and the voltage across it thus varies in exact sympathy with the original audio signal.

The audio across C_2 is governed in amplitude by the extent of the change in width of the anode pulses and in frequency by the rate of change of the width.

It should be noted that on 625 lines switch S_1 (Fig. 3.13) closes. This shorts out the 405 audio load resistor R_1 and connects the secondary of the f.m. final sound i.f. transformer T_1 between grid 1 of V_1 and chassis. S_2 opens and starts the oscillator on grid 3. The bias is also changed from that required for a straight a.f. amplifier to that for short grid base operation of the valve. This circuit has several advantages over the ratio detector. It has good limiting action. The ratio detector transformer is replaced by a simple tuned circuit. A pair of diodes and a triode valve or valve section are also saved, for the audio output is of sufficient level to drive a pentode output valve direct.

De-emphasis is provided by the time constant formed by the anode load R_5 and the integrating capacitor C_2.

Small 38·15Mc/s Losses

Another method of accommodating both 38·15Mc/s 405 sound and 6Mc/s 625 intercarrier sound inputs is shown in Fig. 3.14. On 405 lines, S_1 is open and the intercarrier signal feed is short circuited at the take-off point as before. The 38·15Mc/s a.m. sound signal is applied to and developed across T_2, from whence it is fed to the control grid of the i.f. amplifier valve V_1, through the low impedance of T_1 secondary winding and capacitor.

S_2 closes and connects the top end of T_4 primary to chassis through C_2. This action ensures extremely small losses at 38·15Mc/s and can permit the use of a single-valve sound channel (*e.g.* GEC, Sobell and McMichael dual-standard models).

On 625 lines, S_1 closes and S_2 opens. The former puts a short across T_2, *via* C_1 and the latter removes C_2 from T_3/T_4 junction. The 6Mc/s

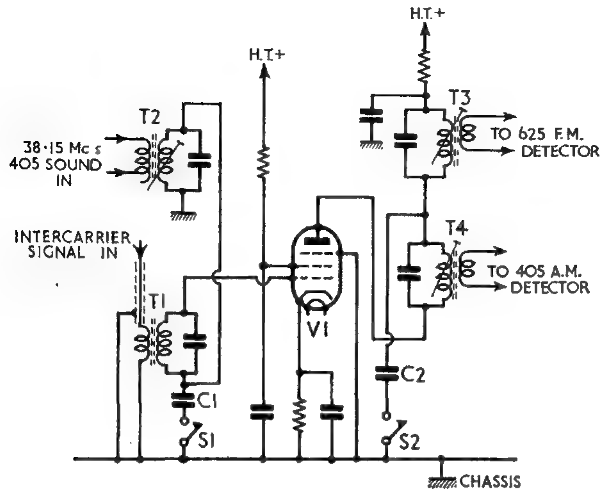


FIG. 3.14. SIMPLE SWITCHING IN THE SOUND CHANNEL AS THIS CIRCUIT SHOWS CAN IMPROVE THE 38·15 MC/S GAIN

intercarrier signal is then allowed to develop across T_3 for feeding the f.m. detector. The series losses of T_4 primary are negligible at 6Mc/s.

Audio Stages

Finally, we come to the audio stages. These differ very little in dual-standard and 625-line-only models from those stages in 405-line-only sets.

A typical (Pye V700D) audio section of a dual-standard set is shown in Fig. 3.15. This employs a triode-pentode valve of the PCL83 (or the more

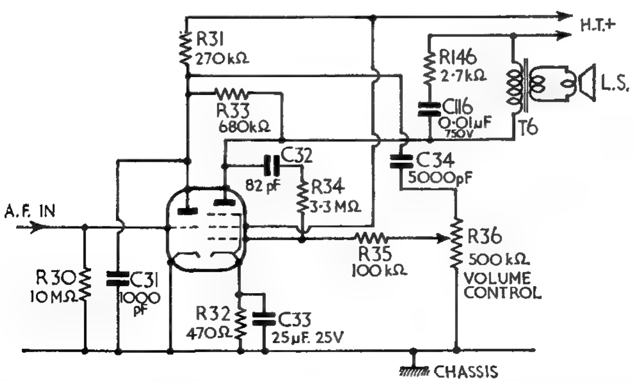


FIG. 3.15. A TYPICAL DUAL-STANDARD AUDIO SECTION USING A TRIODE-PENTODE VALVE. WHEN AN EH90 HEPTODE VALVE IS USED AS THE F.M. DETECTOR/A.F. AMPLIFIER, THE PENTODE SECTION IS USUALLY DRIVEN DIRECT

recent PCL86) variety. The triode is a simple voltage amplifier, accepting the audio from the selected detector at its grid. This drives the pentode output valve, *via* the a.f. coupling capacitor C_{34} and the volume control R_{36}.

Negative feedback is introduced by R_{33}, while frequency-selective feedback is given by C_{32} and R_{34}. Further response correction is provided by R_{146} and C_{116}.

To keep the gain of the triode high, grid current biasing is employed. This is signified by the 10MΩ grid resistor and the absence of a triode cathode resistor. The pentode is biased in the conventional way by the cathode resistor R_{32}. C_{33} in parallel with it avoids degenerative feedback.

On models which feature the EH90 heptode f.m. detector, the triode is often deleted, and the heptode drives a pentode section direct. For example, in the GEC, Sobell and McMichael range of dual-standard sets the EH90 drives the pentode section of a PCL84. The triode section is used in a vision/noise cancelling circuit.

New valves are now being used in dual-standard models, an interesting version being a double-pentode, using a 10-pin Decal base. This has various applications, since the sections are of frame grid construction.

In one application one section is used as the video amplifier and the other as a.g.c. amplifier, sync. separator or intercarrier sound amplifier. The frame grid construction of the video amplifier results in a higher slope and much larger peak anode current than obtainable from the earlier valves used in this service. These valves will be referred to in later sections.

4

SYNCHRONIZING, CONTRAST CONTROL AND A.G.C. CIRCUITS

BECAUSE the picture signal at the anode of the video amplifier valve is negative-going on both 405 and 625 lines, standard change switching is not required in the sync. separator circuits. The majority of models incorporate a conventional pentode sync. separator fed from the video amplifier valve anode circuit in the ordinary manner.

The mean-level type of vision a.g.c. used in a large proportion of 405-line-only sets is also employed in dual-standard models. In Fig. 4.1 is shown such a circuit from models of the K-B VV series. The vision a.g.c. bias is obtained from the control grid circuit of the sync. separator valve, the valve acting in the leaky-grid detector mode.

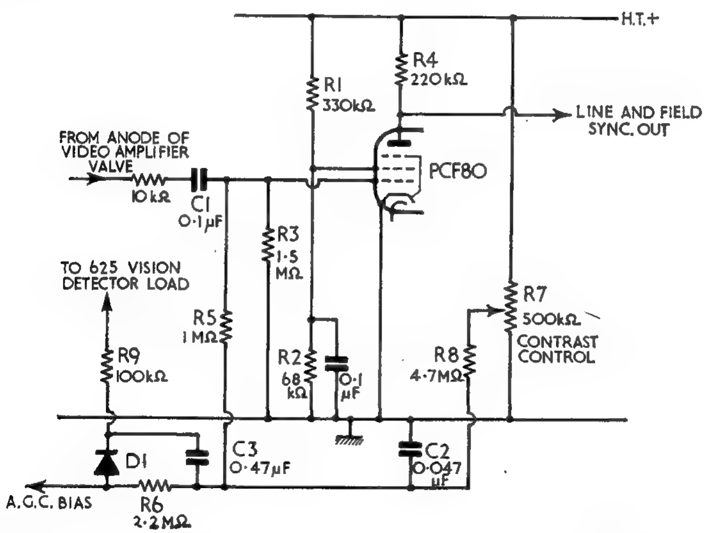


FIG. 4.1. A DUAL-STANDARD SYNC. SEPARATOR STAGE AND MEAN-LEVEL VISION A.G.C. SYSTEM, WITH MANUAL CONTRAST CONTROL

Sync. Separator Action

From the sync. separator aspect, the valve has a short grid base due to the low screen and anode voltage, the screen being fed from the potential-divider R_1 R_2. The valve is pushed into grid current on the positive-going tips of the sync. pulses and into anode current cut-off on the negative-going picture signal.

During the sync. pulses grid current flows in R_3 and charges capacitor C_1. The grid thus becomes negative with respect to the cathode (chassis).

During the picture part of the signal C_1 tends to discharge through R_3. The valve is effectively biased by the signal, therefore, and the voltage at the grid is somewhat influenced by the nature of the picture signal.

This leaky-grid bias places the valve towards anode current cut-off, and anode current flows essentially during the time of the sync. pulses, while during the time of the picture signal the anode current is zero. The sync. pulses thus form across the anode load R_4, and these are conveyed to the field and line timebases through the appropriate networks.

When the valve is passing grid current on the tips of the sync. pulses, the impedance between the control grid and cathode falls to a low value. This has the effect of clipping the tips of the pulses and ridding them of noise and interference.

The amount of grid current flowing in R_3 is proportional to the amplitude of the video signal applied to the sync. separator valve. The larger the signal, the greater the grid current and the greater the voltage drop across R_3.

Mean-Level A.G.C.

Since the voltage at the grid is negative with respect to chassis it is suitable for use as an a.g.c. bias, and is so exploited in many single and dual-standard sets. With a rise in signal strength the bias applied to the controlled valves increases negatively, thereby reducing their effective mutual conductance and the gain of the stages, and adjusting the signal level to its original pre-established value. The converse happens, of course, when the signal strength falls.

In Fig. 4.1 the negative sync. separator potential is applied to the a.g.c. line through the "hold-off" resistor R_5. Filtering is provided by C_2, and additionally by C_3 and R_6. There may also be extra filtering along the line.

On 405-line-only models, diode D_1 is connected between the a.g.c. line and chassis to prevent the a.g.c. line from going positive, which could happen due to a fault condition or in the event of failure of the vision transmission. Should the a.g.c. line tend to go positive the diode would conduct and effectively short the positive potential to chassis.

On dual-standard models the diode still acts in this way on 405 lines, but serves a slightly different purpose on 625 lines, as we shall see.

Manual Contrast Control

When vision a.g.c. of any kind is adopted, the contrast control can no longer be the ordinary type of r.f. gain control, but it must be in some way related to the a.g.c. function, so as to set the a.g.c. **take over** level. An r.f. gain type of preset may be used simply to adjust the "sensitivity" of the r.f. amplifier, but in most modern sets this is now abandoned in favour of some automatic control.

A typical manual contrast control system incorporates a potentiometer across the h.t. supply, the slider of which is connected to the a.g.c. line. Such is used in the circuit of Fig. 4.1.

R_7 is the contrast control and R_8 is the "hold-off" resistor, to keep the impedance of the a.g.c. line high and of fairly constant value over the

range of the contrast adjustment. With the slider of R_7 at the "earthy" end of the control, the full negative voltage as produced at the sync. separator grid, and as governed by the circuit elements, is applied to the a.g.c. line. The contrast is then at minimum (minimum gain), since the controlled valves are fully biased negatively.

When the control is turned in the opposite direction, a positive potential from the h.t. line is reflected on to the a.g.c. line, the current being limited and the impedance held high by R_8. This tends to counteract the negative bias from the sync. separator and reduce the negative bias on the controlled valves. The contrast is then at maximum (maximum gain).

The control allows variations of contrast (vision channel gain) between minimum and maximum.

Varying levels of a.g.c. and manual bias are usually applied to the vision i.f. amplifier and r.f. amplifier valves in a "sequential" fashion. That is, bias is applied first to the controlled i.f. valve or valves and then, after the signal reaches a higher level, to the r.f. valve. An arrangement of biased diodes is used for this application.

The idea is to maintain the best possible signal/noise ratio at low signal levels while avoiding overloading and intermodulation at high signal levels. The former would not be possible if the r.f. amplifier valve were biased negatively (apart, of course, from its standing bias) on weak aerial signals, even though the resulting bias may only be small. Thus, the bias to the r.f. valve in the tuner is held off until the input signal rises above a certain level.

Although the vision a.g.c. system as detailed above is termed **mean-level,** there is really no mean vision signal level with the positive modulation of the 405-line signal, since the level of modulation rises with the brightness of the transmitted scene. This can detract a little from the picture quality, but it remains domestically acceptable when mean-level a.g.c. is used.

So far we have considered the sync. separator and vision a.g.c. mainly on the 405-line standard. On 625 lines, the sync. separating process is little altered, but in the provision of a.g.c. there is an extra problem.

Blocking on 625 Lines

This relates to a phenomenon called **blocking.** The effect applies mainly to mean-level a.g.c. systems which need to suit both standards. In an overload condition the sync. pulses of the signal can be severely suppressed, if not cut off completely, at the grid of the video amplifier due to the negative modulation of the 625 signal.

This is reflected back to the sync. separator valve so that no negative voltage is produced at the control grid, resulting in a loss of a.g.c. action and an overload. The set is then caused to work at maximum gain, contrast control adjustment failing to remedy the condition.

In practice, the effect is avoided by applying to the vision a.g.c. line a small negative voltage as obtained from the vision detector when this is switched for 625-line operation. Resistor R_9 in Fig. 4.1 is for this purpose.

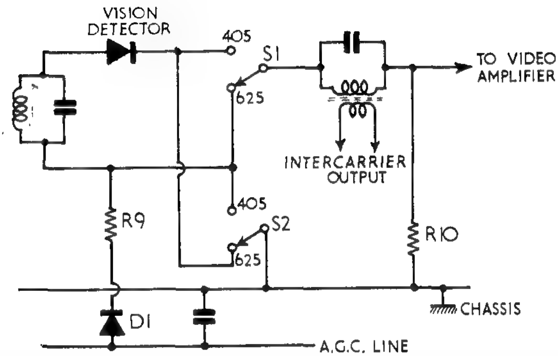


FIG. 4.2. ON 405 LINES DIODE *D1* IS CONNECTED BETWEEN THE CHASSIS AND A.G.C. LINE AS A CLAMP TO PREVENT THE LINE FROM RISING POSITIVELY DUE TO THE POSITIVE VOLTAGE REFLECTED FROM THE CONTRAST CONTROL (SEE FIG. 4.1). ON 625 LINES THE DIODE IS CONNECTED TO THE "ANODE" OF THE VISION DETECTOR DIODE AND THUS PASSES A LITTLE OF THE D.C. COMPONENT OF THE VISION SIGNAL TO THE A.G.C. LINE, THROUGH *R9*, TO AVOID BLOCKING DUE TO THE NEGATIVE MODULATION CAUSING OVERLOADING AT THE VIDEO AMPLIFIER VALVE GRID. SUCH OVERLOADING CAN CUT OFF THE SYNC. PULSES AND THE NEGATIVE A.G.C. BIAS AT THE SYNC. SEPARATOR GRID

Fig. 4.2 shows the circuit section in greater detail. Here we have the vision detector, with R_9 and D_1 connected to the a.g.c. line. On 405 lines S_2 puts D_1 down to chassis through R_9 so that the diode acts as a "clamp", as already explained.

On 625 lines R_9 and D_1 are effectively connected to the "hot" end of the detector load R_{10}, the negative voltage present at that point being passed through R_9 and D_1 to the a.g.c. line. This avoids the a.g.c. from locking on to zero voltage while always keeping the line a little negative to prevent the blocking trouble.

While a satisfactory "black-level" a.g.c. system is not possible using just a single stage on the positive modulation of the 405 system, a virtual black level system is feasible, using a single valve function, on the negative modulation of the 625 system. Note here that "black-level" refers to a system where the gain of the vision channel is maintained independent of picture content of the signal.

Black-Level Vision A.G.C.

Many 405-line-only sets incorporate a black level vision a.g.c. system, the idea being to "sample" the black level of the signal either on the porches of the line sync. pulses or during the field pulse period by means of a circuit which is "gated" either by the line (Pye, etc.) or field (Murphy) timebases. The signal at that black-level instant is then rectified and used as an a.g.c. bias.

On the negative modulation of the 625-line system a simple gated black level circuit can be employed with excellent results. The idea is to provide d.c. coupling from the video amplifier anode to an a.g.c. valve. This holds the **tips** of the sync. pulses at a constant level and ensures that the gain is controlled independently of the picture content of the composite video signal.

Mixed A.G.C.

A circuit has been described* which utilizes mean level a.g.c. on 405 lines and black level a.g.c. on 625 lines by simple switching. The circuit is shown in Fig. 4.3.

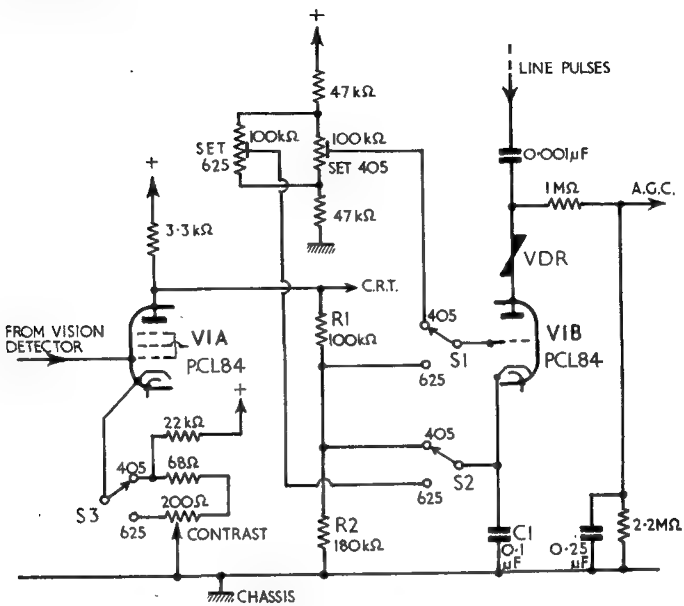


FIG. 4.3. VISION A.G.C. CIRCUIT WHICH PROVIDES MEAN-LEVEL A.G.C. ON 405 LINES AND BLACK-LEVEL A.G.C. ON 625 LINES

Here V_{1A} is the video amplifier valve and V_{1B} the a.g.c. valve. V_{1B} grid is d.c. coupled to V_{1A} anode on 625 lines through S_1 while the cathode is set at a potential as governed by the "set 625" preset, sufficient to render the valve just conducting on the tips of the positive-going sync. pulses.

V_{1B} is, in fact, a form of pulse amplifier and is operated by pulses applied to the anode circuit from the line timebase. The voltage-dependent resistor (VDR) rectifies the pulses and the valve works as a controlled d.c. restoring circuit. The potential developed across the VDR is proportional to the conduction of the valve, the greater the conduction, the greater the voltage produced. This voltage is used as an a.g.c. bias which, of course, varies in sympathy with the amplitude of the sync. pulses.

On 405 lines, the video signal is integrated by R_1, R_2 and C_1, and the resulting mean d.c. potential is applied to V_{1B} cathode *via* S_2. The grid potential is adjusted by the "set 405" preset so that the mean d.c. of the cathode signal varies the valve conduction as its level changes with change

*J. H. Haslett & P. L. Mothersole, "Dual Standard Video, Synchronising and A.G.C. Circuits," *Journal Brit. I.R.E.*, November, 1962.

in signal strength. This in turn varies the voltage at the valve anode and thus provides a suitable a.g.c. bias.

Manual control of contrast is achieved by varying the working point of the video amplifier valve by means of a low value potentiometer in its cathode circuit. By varying the working point, the anode voltage is varied and this is reflected into the a.g.c. circuit as a change in a.g.c. bias.

Switch S_3 in V_{1A} cathode simply reverses the connections to the contrast control over the two standards, for unless this is done the direction of rotation of the control for increasing the picture contrast will not be the same on each standard owing to the different polarities of the vision modulation.

Although very few commercial receivers employ mixed a.g.c. systems, as described, these are quite feasible, as also are adaptions of 405-line-only black level and gated circuits for use on both standards.

Manual Contrast

We have so far seen that manual control of picture contrast in dual-standard models is provided by utilizing the a.g.c. delay potential, as in mean level 405-line-only sets and by arranging for the operating point of the video amplifier valve to be varied.

In this latter respect, the screen grid potential, instead of the cathode, can be varied by the control. With this method the black level of the video signal remains almost constant.

The control can be made automatic by the use of a cadmium sulphide cell, such as the Mullard ORP12, connected into the screen grid circuit in such a way that the screen potential is caused to rise as the ambient illumination falling upon the cell rises, thereby giving an increase in contrast.

Alternatively, the contrast control is sometimes connected in the anode circuit of the video amplifier, as shown in Fig. 4.4. The control may be "equalized" over the video spectrum and it then works rather like the optical counterpart of the hi-fi "loudness control".

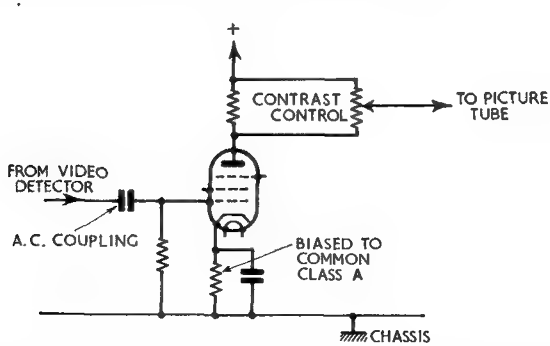


FIG. 4.4 METHOD OF MANUAL CONTRAST ADJUSTMENT BY THE USE OF A POTENTIOMETER IN THE ANODE CIRCUIT OF THE VIDEO AMPLIFER VALVE. THIS HAS THE ADVANTAGE THAT THE SIGNAL LEVEL AT THE VISION DETECTOR REMAINS CONSTANT, THEREBY AVOIDING UNDUE INTERFERENCE WITH THE INTERCARRIER SOUND SYSTEM

Remote Control of Contrast

Equalization is needed to counteract the capacitive load that the control adds to the anode circuit of the video amplifier valve. A suggestion which has been put forward so that the contrast control may be isolated completely from the signal circuits is that a cadmium cell be arranged in a potential-divider network instead of the ordinary contrast control and that the cell be illuminated by a small lamp. As the resistance of the cell is dependent upon the illumination falling upon it (its resistance falls with increasing illumination), the control of contrast can be achieved by arranging for the control to vary the illumination of the lamp!

Another point in favour of having the contrast control in the anode circuit of the video amplifier is that the signal amplitude is unaffected at the vision detector. So far as the intercarrier sound function is concerned this is an important point, as will now be appreciated.

The use of the new "double-pentode" valve, with one frame grid pentode acting as the video amplifier facilitates video valve anode contrast control, since the valve slope is that much higher and the valve is capable of much larger peak anode currents than earlier video amplifiers.

On a negative-going signal for the 625-line pictures a pulse of interference rises in a positive direction, as shown in Fig. 4.5. The pulse is thus rising towards black level, meaning that instead of the white

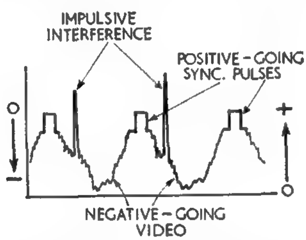


FIG. 4.5. SHOWING HOW IMPULSIVE INTERFERENCE RISES POSITIVELY ON NEGATIVE-GOING PICTURE SIGNAL (*Mullard*)

interference spot associated with the 405-line system, the pulse causes a grey or black spot on the 625-line standard. From this aspect, therefore, a vision interference limiter of the kind used in 405-line-only sets is not required.

Such a limiter is needed, though, on 405 lines, and a white spot inverter or "spotter" is often employed, this being switched out on 625 lines.

However, from Fig. 4.5 it is seen that the interference pulses could be mistaken by the sync. separator for sync. pulses. This, indeed, can happen and impairment to both the line and the field synchronizing can result, especially in areas of weak signal and heavy interference.

On the line the trouble is avoided by the use of flywheel sync. on the 625-line standard, for apart from interference of a general nature ragged verticals can result from h.f. noise in the receiver itself. This is more noticeable, incidentally, at the faster scanning speed.

On the field timebase random judder or intermittent rolling can prove troublesome in the event of heavy impulsive interference. This is avoided

by the use of interference cancellation circuits now to be described. Note, however, that external impulsive interference is far less of a bother on the u.h.f. bands than on Bands I and III. It will have already been noticed in practice that on the v.h.f. bands such interference drops considerably on Band III as compared with Band I. A further drop equally as beneficial occurs from Band III to Bands IV and V. Nevertheless, interference cancellation circuits are still desirable.

Fig. 4.6 shows the basic circuits of a dual-standard video amplifier, interference suppressor, interference inverter or cancellation artifice and sync. separator, at valves V_1, V_2, V_3 and V_4 respectively.

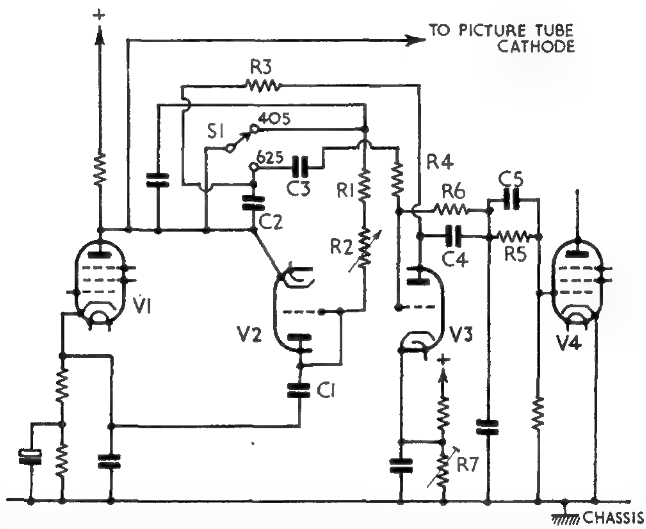


FIG. 4.6. IN THIS CIRCUIT *V1* IS THE VIDEO AMPLIFIER, *V2* THE 405-LINE INTERFERENCE LIMITER OR SPOTTER, *V3* THE INTERFERENCE CANCELLATION STAGE AND *V4* THE SYNC SEPARATOR. ON 405 LINES *V2* IS IN CIRCUIT AND *V3* MUTED WHILE ON 625 LINES *V3* IS IN CIRCUIT AND *V2* MUTED, THIS BEING ACCOMPLISHED BY *S1*

Dual-Standard Limiter

On 405 lines V_2 in series with C_1 is connected between the anode and cathode of the video amplifier valve V_1. V_2 is connected as a diode with its anode and grid joined. The valve becomes conductive when its cathode goes negative with respect to its anode.

Under zero interference conditions C_1 charges and the valve V_2 is non-conducting. During an interference pulse, however, the cathode goes negative, the valve conducts and short circuits the interference pulse, and C_1 discharges through R_1 and R_2, with R_2 being variable to adjust the C_1 R_1+R_2 time constant, and thus the limiting conditions in the ordinary manner.

On 625 lines, S_1 changes over and h.t. is removed from V_2 anode and the valve is muted. H.T. is, however, applied to the anode of V_3 through

the switch and R_3, this being isolated from h.t. on 405 lines by C_2. R_3 then forms the anode load of V_3.

Signal from the anode of the video amplifier valve V_1 is applied to the grid of V_3, *via* S_1 and the network C_3 and R_4. Sync. pulses reach the control grid of the sync. separator valve V_4 by way of C_2, R_3, C_4 and the parallel combination C_5 R_5.

Now, V_3 conduction is set by the preset R_7 so that under interference-free conditions the valve is non-conducting. However, in the event of an interference pulse reaching the sync. separator V_4 grid *via* the route detailed above, the pulse will also be applied to the grid of V_3, *via* R_6, which is connected to the junction of C_4 R_5.

The pulse causes V_3 to conduct. The pulse is thus amplified by the valve and appears in inverse phase at the anode. It is then reinserted in the same route as before, *via* C_4, and since there are now two pulses of opposite phase in the circuit at the same time complete cancellation occurs and no interference pulse arrives at the sync. separator grid.

Switch S_1 thus changes from the ordinary 405-line white spot suppressor to the interference pulse cancellation circuit.

It is interesting at this juncture to consider the operation of a conventional pentode sync. separator stage (Fig. 4.1) under conditions of impulsive interference. It will be recalled that under normal signal conditions the sync. pulses themselves push the stage into grid current and establish the clipping level of the circuit so far as the picture content is concerned.

In the presence of strongly positive-going interference pulses, however, the stage is pushed much harder than normal into grid current, thereby charging the control grid capacitor to a value in excess of its nominal charge due to the sync. pulses.

This abnormally high charge holds the stage at anode current cut-off, even during the sync. pulses, for a period determined by the control grid time constant (C_1, R_3, Fig. 4.1), as shown by waveform (a) (Fig. 4.7). The

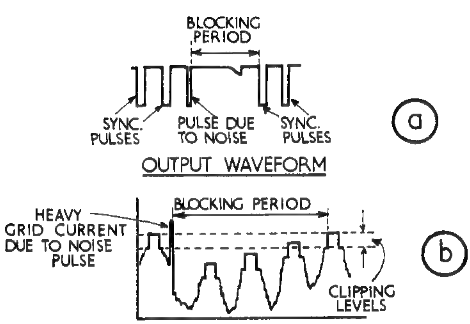


FIG. 4.7. THE WAVEFORMS CAUSED WHEN THE SYNC. SEPARATOR "BLOCKS" DUE TO IMPULSIVE INTERFERENCE ON 625 LINES, AS DESCRIBED IN THE TEXT (*Mullard*)

period of cut-off is called the **blocking period** and is illustrated by waveform (b). This is the output waveform, of course, and here a

spurious pulse is created due to the interference pulse, after which there is virtually zero output until the blocking charge is exhausted.

Noise-Gated Sync. Separator

As this phenomenon is peculiar to the sync. separate stage proper, no amount of juggling after the stage will correct it. The only solutions lie in the "noise inverter" circuit, previously described, or in a "noise-gated" sync. separator stage.

This latter device provides for interference cancellation in the sync. separator itself, and a circuit adopting the ECH84 triode-heptode valve has been evolved by Mullards. The ECH84 has been designed specifically to supersede the ECH81, sometimes used in 405-line-only sync. separator circuits. The ECH84 will now be found in dual-standard models.

The heptode section of the valve is arranged as the sync. separator proper while the triode section is usually given over to pulse clipping. The basic circuit, with waveforms, is shown in Fig. 4.8.

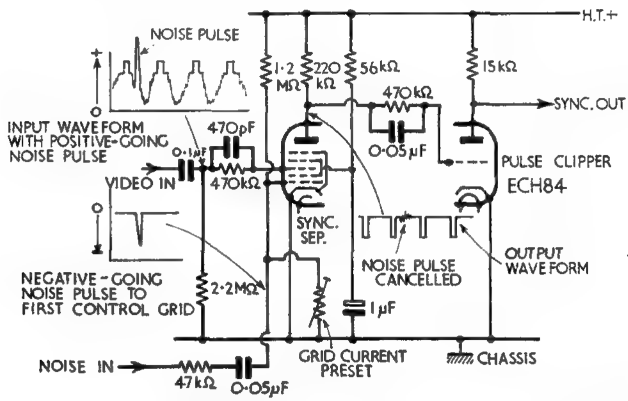


FIG. 4.8. IN THIS HEPTODE SYNC. SEPARATOR CIRCUIT A NEGATIVE-GOING INTERFERENCE PULSE APPLIED TO THE FIRST CONTROL GRID CANCELS THE POSITIVE-GOING INTERFERENCE PULSE WHICH IS PRESENT ON THE SECOND CONTROL GRID

Note that the second control grid of the heptode is used like the control grid of an ordinary pentode sync. separator, this being coupled through the time constant network to the anode of the video amplifier valve.

To the first control grid, however, noise signals only are applied. In addition to these—which are now negative-going—the grid is given a positive potential so that a predetermined amount of grid current flows. When the first control grid receives a negative-going noise pulse the valve is cut-off and the positive-going version of the same pulse at the second control grid, along with the vision signal, is thereby deleted from the anode circuit. In other words, the negative-going pulse at the first control grid acts as a gating signal to cut off the valve for the duration of the pulse.

The result is almost complete cancellation of the noise pulse at the anode of the valve, as shown by the output waveform in Fig. 4.8. The

effect is rather the same as though the pulses were "punched" out of the sync. waveform.

The output waveform is then applied to the grid of the triode section of the valve, this serving as a pulse clipper and the output being developed across the anode circuit.

In a practical circuit there is usually a preset control for adjusting the grid current in the first control grid. This setting is rather critical, as the positive potential applied to the grid must be large enough to maintain grid current during normal video signals, but not so great that negative-going noise pulses will fail to pull the valve out of grid current.

As these conditions are somewhat related to the setting of the contrast control, the circuit may appear as shown in Fig. 4.9. Here the positive

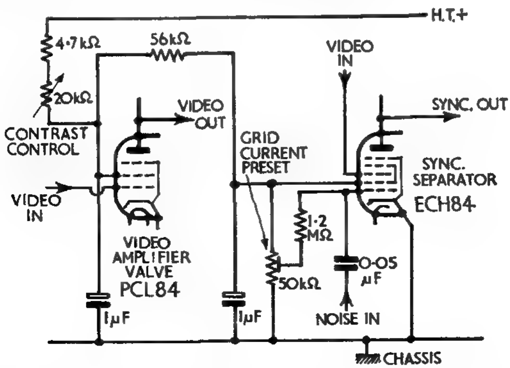


FIG. 4.9. TO OPTIMIZE INTERFERENCE CANCELLATION AT ALL CONTRAST SETTINGS, THE GRID CURRENT IN GRID 1 CIRCUIT IS MADE DEPENDENT UPON THE CONTRAST CONTROL SETTING

delay is varied in step with the amplitude of the video signal, the contrast control, in fact, altering the conditions of the video amplifier, as we have already seen.

The negative-going noise pulses can either be obtained from the output of the vision detector or from a separate and independent "noise detector". The latter arrangement is the less critical and most efficient, and where the noise signal can be made completely free from video information a grid current preset may not be used.

The noise pulses must be separated from the sync. pulses, and one way of accomplishing this is by the use of a frequency selective network, having in mind that noise pulses are composed of signal components from very low to very high frequencies, while the signal components of the sync. pulses are contained in a relatively narrow channel of about 1Mc/s, centred on the vision carrier.

The circuit of a frequency selective noise detector is shown in Fig. 4.10. The tuned *LC* circuit is adjusted for **minimum** low frequency signal at the noise detector output, an action which provides the required discrimination between the sync. and noise and successfully eliminates the sync. pulses whilst retaining the noise pulses.

The triode section of the valve is sometimes used as a noise detector, as shown in Fig. 4.11. Here a noise detector transformer couples the

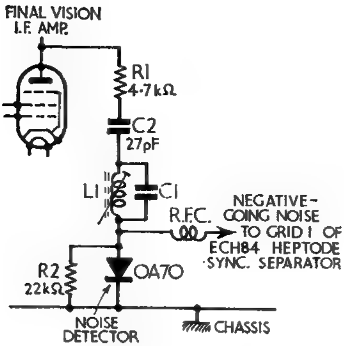


FIG. 4.10. SHOWING HOW THE NOISE PULSES ARE PRODUCED BY A FREQUENCY-SELECTIVE NOISE DETECTOR. THIS PASSES THE NOISE PULSES YET ELIMINATES THE SYNC. PULSES

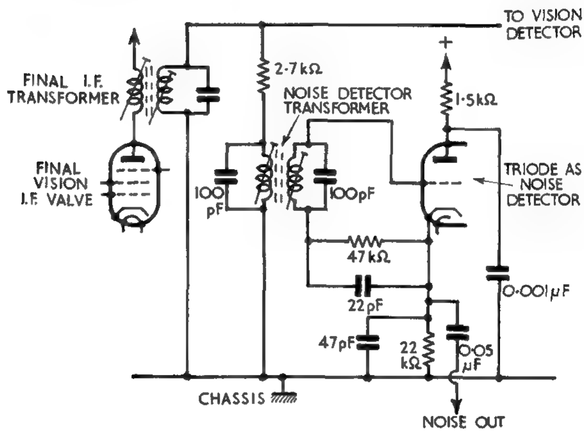


FIG. 4.11. THE TRIODE OF THE ECH84 CAN BE USED AS A NOISE DETECTOR, AS THIS CIRCUIT SHOWS

signal from the final vision i.f. amplifier to the triode grid, the passband being about 1·5Mc/s, centred about 3·5Mc/s. The triode is arranged after the style of a leaky-grid detector and the negative-going noise output is developed across the cathode resistor.

To avoid random picture jump or roll under conditions of extremely heavy interference, when cancellation could occur on a sync. pulse, the field sync. coupling is kept as light as possible so that the synchronized field frequency is close to the timebase free-running speed. Alternatively, the field sync. pulses may be shaped so that the free-running speed of the field timebase is exactly the same as the synchronized speed.

On line the trouble does not arise on 625 lines since flywheel line sync. is invariably adopted.

The sync. pulses are, in general, extracted from the sync. separator stage and applied to the field and line timebases in the usual manner through integrating and differentiating circuits respectively.

5

TIMEBASE CIRCUITS

A PROPOSITION was put forward at one time in Great Britain which would have involved the use of a 60c/s field frequency—in conformity with the American standard—but was not taken up. On both the British 405-line and 625-line systems a 50c/s field frequency is adopted.

Probably the chief reason for this is that 50c/s matches the mains power distribution frequency, meaning that the field frequency can, in many instances, be "locked" to the power frequency. Such synchronizing is desirable to avoid a slow beat ripple effect on the picture which can result from the use of a field frequency which differs a few cycles or so from the power frequency.

Asynchronous Working

The ripple is caused by residual hum in the picture circuits of a receiver. Under normal conditions, when the field is locked to the power frequency, the effect of the hum is insignificant. However, when the hum causes a low-frequency beat effect, compression and expansion of the picture occurs at the beat frequency producing the picture ripple, as mentioned.

It can be cleared, of course, by eliminating all traces of hum from the vision stages of the set; but that can be an added cost. Nevertheless, it would seem that steps will be taken by set makers to reduce, at least, the residual hum content, for there are possibilities that later "asynchronous" working (where the field frequency at the transmitter is not purposely locked to the power frequency) will be adopted by transmitting authorities, especially when colour is launched. Already tests are being conducted with asynchronous transmission in Great Britain.

This, then, means that switching in the field circuits is not required in British dual-standard receivers.

What is necessary, of course, is a change in line frequency. A complete picture on both standards is composed of two interlaced fields. On 405 lines each field is made of $202\frac{1}{2}$ lines. Thus, with a field frequency of 50c/s the 405-line line timebase needs to work at $202\frac{1}{2} \times 50$ c/s, which is 10,125c/s. On 625 lines the line timebase needs to work at $312\frac{1}{2} \times 50$ c/s, which is 15,625c/s. Aside from all the other factors examined in past chapters, therefore, the standard change switch needs to change the line timebase from 10,125c/s on 405 lines to 15,625c/s on 625 lines.

While the actual change in line frequency is not too difficult to accomplish, there are several by-product problems which are created by this change, as we shall see.

Need for Flywheel Control

The line sync. is affected by interference and random noise on both standards, and on many 405-line-only receivers flywheel sync. is employed to combat the disturbance. On 625 lines, however, the disturbance is far

more noticeable mainly due to the higher scanning speed and energy. As this cannot be eliminated by direct synchronizing, flywheel line sync. is imperative on the 625-line standard. The use of a simple flywheel controlled oscillator solves the problem adequately on most receivers.

Flywheel sync. may be used on both standards on some models, though switching from direct to flywheel sync. may be utilized on others. Convertible-to-switchable models may incorporate direct sync. when the set is arranged for 405-line-only reception and flywheel sync. after conversion.

Flywheel controlled line oscillators for dual-standard applications are mainly of the multivibrator or sine wave type. The latter, for example, is used in some of the early 405-line-only sets, such as in Ferguson and GEC and the former adopted in some Pye models.

Pye Circuit

The circuit of the dual-standard Pye line oscillator is shown in Fig. 5.1. Valve sections V_2 and V_3 are wired in the form of a cathode-coupled

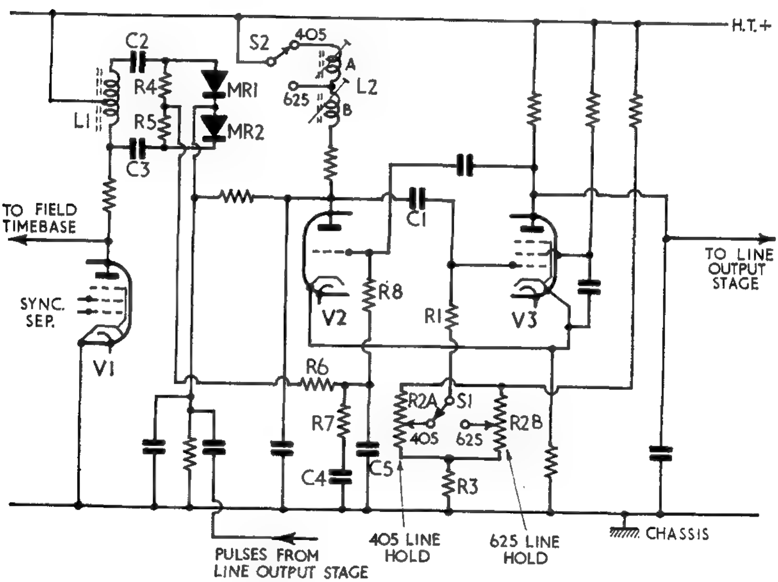


FIG. 5.1. A DUAL-STANDARD FLYWHEEL CONTROLLED MULTIVIBRATOR LINE OSCILLATOR BY PYE

multivibrator. The frequency of oscillation is governed by the time constant formed by C_1 in series with R_1, R_2 and R_3. The frequency is adjustable by the line hold control R_2.

Apart from the normal anode resistor, the anode of the sync. separator V_1 is loaded by a small inductor L_1. Across this the line sync. pulses are developed and are applied to two metal rectifiers, MR_1 and MR_2, through

C_2 and C_3. The metal rectifier and inductor circuit, including R_4 and R_5, represents a discriminator.

In addition to the line sync. pulses, this has fed to it pulses from the line output stage. Accurate synchronizing of the system is achieved when the frequency and phase of the line output stage pulses match those of the sync. pulses. This is because, in effect, the voltage at the junction of R_4,R_5 is zero under synchronous conditions.

However, should the frequency of the line oscillator tend to wander, the discriminator balance would be disturbed. A control potential would then develop at the junction of R_4,R_5, the magnitude and polarity of which would depend upon the phase difference between the sync. and line pulses and whether the oscillator signal is tending to lead or lag the sync. signal.

It will be seen that the control potential so derived is fed to the grid of V_2 through R_6 and R_7. This serves to control the frequency of the multivibrator and thus bring it back into step automatically with the sync. pulses. So far as the line oscillator itself is concerned, the effect is the same as if the line hold control were adjusted manually. This control, of course, is adjusted in the first place so that the line frequency is brought within the "pull-in" range of the discriminator.

R_7, C_4 and C_5 are incorporated to avoid "hunting" effects. A fairly long time constant is also put on the control line by C_5 to assist with the flywheel effect.

Note that on the "405-line" position R_{2A} is the line hold control while on the "625-line" position R_{2B} is the line hold control. The reason for the two controls, switched by S_1, is so that the line speed can be established on both standards initially. Readjustment of the line hold control is then avoided when a viewer changes from one standard to another.

On some models R_3 may be a preset control so that the main line hold controls can be arranged to give optimum lock towards the centre of their ranges. Other models may feature only one line hold control, but then there is a preset control which can be adjusted to ensure that the locking point on the main control is the same for both standards (see Fig. 5.2).

The flywheel effect in the circuit of Fig. 5.1 is provided chiefly by L_2 in the anode circuit of V_2. On 405 lines sections (A) and (B) in series are adjusted to resonate to 10,125c/s, while on 625 lines S_2 by-passes section (A), and section (B) only is in circuit and is adjusted to resonate to 15,625c/s. The idea is first to adjust the core in section (B) on 625 lines and then the core in section (A) (leaving core (B) alone) on 405 lines.

The oscillatory effect created in the anode circuit of the multivibrator tends to hold the line sync. for brief periods even though the line sync. pulses may be missing or badly affected by impulsive interference.

The circuit averages the sync. pulses over a number of lines. This means that random noise cannot cause the displacement of individual lines. The worst that can happen is slight bending of the verticals due to the production of a false control voltage at the discriminator output, as may result from noise, interference or spurious signals.

On a set without flywheel sync. under such conditions the vertical parts of a picture would tend to be very ragged, giving a sawtooth effect, and in severe cases whole slices of the picture may be torn away from the main picture, especially on 625 lines.

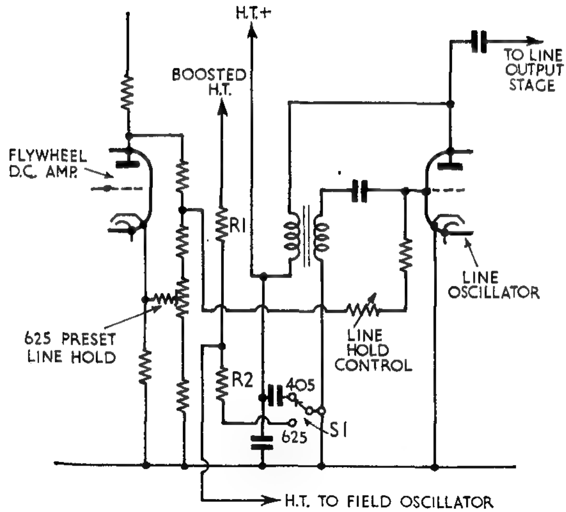


FIG. 5.2. IN THIS THORN SERIES CIRCUIT A BLOCKING OSCILLATOR TYPE OF DUAL-STANDARD OSCILLATOR IS USED. CONTROL IS FROM A DISCRIMINATOR VIA THE D.C. AMPLIFIER SHOWN. A PRESET ADJUSTMENT IS AVAILABLE FOR BALANCING THE LINE LOCK ON 625 LINES TO THE POSITION OF 405 LOCK ON THE MAIN CONTROL. *S1* ADJUSTS THE TIME CONSTANT CAPACITOR AND THE H.T. SUPPLY TO THE FIELD OSCILLATOR (SEE TEXT)

Sine Wave System

A typical flywheel controlled sine wave line oscillator is shown in Fig. 5.3. There are three stages. The discriminator employing two diodes, as in the previously considered circuit; the line oscillator V_2; and a reactance valve V_1.

The reactance valve reflects a capacitance effect across the tuned oscillator circuit (L_1 and C_1—or C_2/C_3), the value of the capacitance being governed by the effective mutual conductance (g_m) of V_1. Thus, by varying the mutual conductance of the valve, altering its bias for instance, the line frequency is caused to alter.

The nominal 625 line frequency is established by L_1 and C_1, since S_1 is open on 625 lines. Fine adjustment of the frequency is possible by the line hold control as this constitutes a potentiometer across the h.t. supply, the slider being taken to V_1 grid circuit (as a means of adjusting V_1 bias and its capacitance as reflector across the oscillatory circuit).

Now, when the sync. pulses applied to the discriminator are in step with the reference pulses fed in from the line amplifier, there is zero control voltage applied to V_1 grid from the discriminator. However, should the line oscillator tend to wander a control voltage is produced which alters the bias on V_1 and in turn changes the capacitance across V_2 in such a way that the correct frequency is restored. The flywheel effect is given by the long time constant of the control line from the discriminator to the reactance valve grid.

On 405 lines S_1 closes and puts C_2 and C_3 in parallel with C_1. This reduces the frequency from 15,625c/s to 10,125c/s. C_3, being a preset

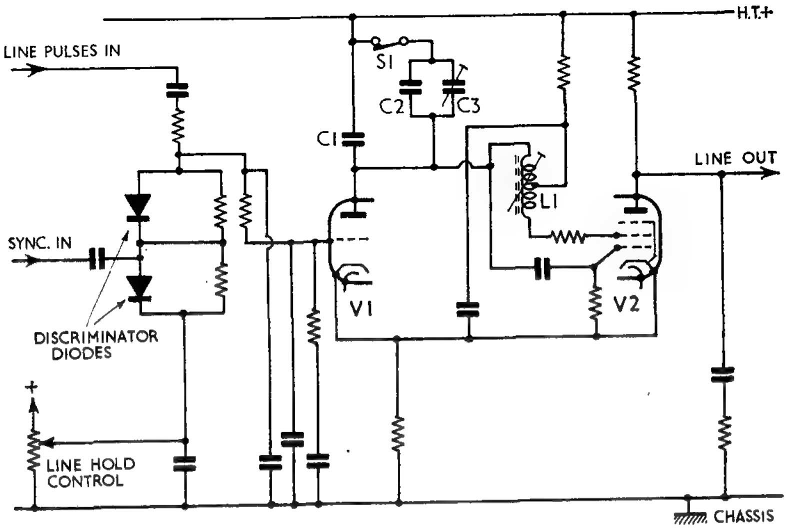


FIG. 5.3. A SINE WAVE TYPE DUAL-STANDARD FLYWHEEL CONTROLLED LINE OSCILLATOR. THE COUPLING BETWEEN THE OSCILLATOR AND THE LINE OUTPUT STAGE GIVES THE CORRECT WAVEFORM FOR WORKING THE OUTPUT STAGE

trimmer, can be used to balance the lock on 405 lines so that it occurs at the same position on the main line hold control as does the 625-line lock.

In Fig. 5.2 the line oscillator is in the form of a blocking oscillator. This is controlled from a d.c. amplifier which itself is operated from the potential developed in a discriminator circuit of the style of those already considered.

Note in Fig. 5.2 that a preset 625 line hold control is featured. The 405 line lock is first established on the main control and then, without altering the position of the main control, the preset is adjusted for optimum lock on 625 lines. This technique ensures that correct lock occurs on both standards at the same position on the main control.

On some models a separate change-over switch may be employed in the oscillator to change the value of the charging capacitor, as shown in Fig. 5.4.

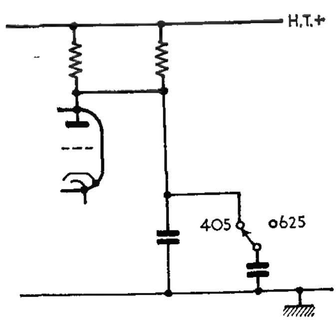


FIG. 5.4. ON SOME CIRCUITS A SEPARATE SWITCH SECTION IS USED TO CHANGE THE CHARGING CAPACITOR AT THE OUTPUT OF THE OSCILLATOR

A line output stage designed for optimum operation on one line standard would fail to operate efficiently on a different line standard. That is, shortcomings in picture geometry, width and brightness would occur if only the repetition frequency of the line oscillator were changed by the "standard change" switch. In practice, therefore, several switching functions are necessary in the line amplifier circuit.

The line output stage is probably the most critical circuit in the whole of the receiver, and even in 405-line-only receivers several artifices have been adopted in the past to improve the overall efficiency. These include the use of low-loss core materials for the line output transformer, more efficient valves, special booster circuits and third harmonic tuning of the transformer. With the employment of flat-faced, wide-angle picture tubes, picture geometry correcting networks are also employed.

All these are set to one line standard, so on changing standard they must all be altered accordingly. In addition, the field oscillator of modern receivers is often energized not from the ordinary h.t. line but from the boosted h.t. derived in the line output stage. As this boost voltage is likely to alter when the line speed is changed, some correction in the field oscillator h.t. supply feed is also necessary.

There is also the possibility of the voltage across the heater of the e.h.t. rectifier valve changing from one standard to another, and this must also be compensated for where necessary.

Dual-Standard Line Output Stage

In Fig. 5.5 is shown a dual-standard line output circuit from sets of the Thorn 850 series. The stage is fairly conventional, apart from the switched components. For instance, C_1 is concerned with the third harmonic tuning of the line output transformer.

Third Harmonic Tuning

On 405 lines, C_{1A} is connected across a part of the transformer to provide the necessary resonance while on 625 lines the resonant frequency is corrected to suit the higher line speed by switch section S_2 switching out C_{1A} and switching in C_{1B}. The tapping position is also changed.

The line scanning coils themselves are energized from the transformer across taps *A-B* on 405 lines and *A-C* on 625 lines, *via* capacitor C_{2A} or C_{2B} respectively, as selected by switch section S_3. The change of tap helps to equalize the width over the two standards while the change in value of C_2 makes the necessary "S correction".

S Correction

Without S correction in some form, modern receivers employing flat-faced, wide-angle picture tubes would display pictures with vertical compression at the centre of the screen and expansion at either side.

The reason for this is that when the electron beam is deflected by the line coils its **length** (from the electron gun in the tube to the fluorescent screen) decreases from a maximum when the scanning spot appears on the extreme left of the screen to a minimum when the spot appears in the centre of the screen. The length of the beam then increases again to a maximum as the scanning deflects the beam from the centre to the right of the screen.

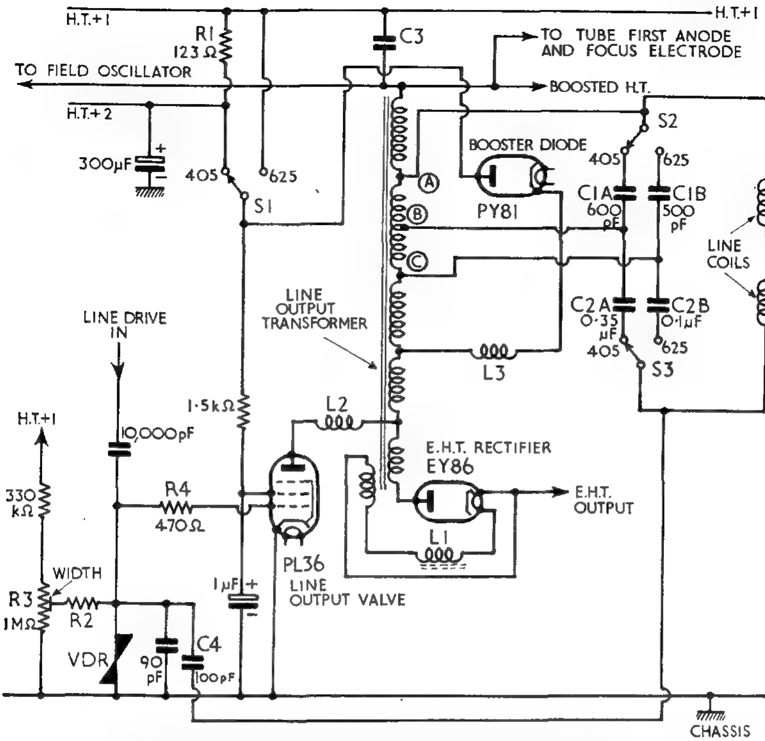


FIG. 5.5. DUAL-STANDARD LINE OUTPUT STAGE. THIS IS A COMPROMISE BETWEEN CONSTANT FLYBACK TIME AND CONSTANT FLYBACK RATIO AND SHOWS ALL THE PARAMETERS THAT NEED TO BE SWITCHED

This results in a changing radius of scan which must be corrected to avoid the distortion mentioned. Correction would not be necessary, of course, if the radius of the screen matched the radius sweep of the beam, for then the electron beam would remain at a constant length over its entire scan.

A capacitor of correct value (to match the line speed) connected in series with the line scanning coils tends to correct the waveform so that the rate of current change in the coils rises towards the centre of the scanning cycle and falls at either side.

With one type of line output stage the voltage across the heater of the e.h.t. rectifier valve tends to rise on changing from 405 to 625 lines. This could lead to premature failure of the valve unless corrected. Switching is virtually impossible in the e.h.t. circuits, of course, but fortunately there is an easy solution to the problem.

Heater Current Correction

The heater is energized by pulse current induced in a few turns of heavy gauge wire wound round the core of the line output transformer.

The current is at line frequency. This means that a frequency selective device can be included in series with the rectifier heater. A capacitance or inductance is frequency selective and in practice an inductance is used, as L_1 in Fig. 5.5.

The number of turns and inductance value given to L_1 are calculated in conjunction with the heater current requirement of the valve. The number of turns of heater winding on the line output transformer is arranged so that the e.h.t. rectifier valve receives the correct heater current through L_1 on 405 lines.

At the higher line frequency of the 625 standard, L_1 offers a greater impedance to the heater current than at the lower frequency of the 405 standard. Thus, the rise in heater current promoted by the higher line speed is neutralized and the heater receives approximately equal current on both standards.

L_1 is just a small coil wound on an iron dust core, the assembly often being embedded in the plastic insulation of the line output transformer close to the e.h.t. rectifier valve holder.

The screen and anode voltages in the circuit of Fig. 5.5 are also decreased when changing from 625 to 405 lines. This is accomplished by switch section S_1 and resistor R_1. Note that R_1 is a dropping resistor from the main h.t. line h.t.+1 to a lower voltage line h.t.+2. The decrease is necessary to give the final "trim" to the stage over the two standards, bearing in mind that the basic efficiency is usually greater on 405 than 625 lines. Hence, the reason for **reducing** the voltage on 405 lines.

R_1 is an h.t. dropping resistor, feeding h.t.+2 line (h.t.+1 is connected to the smoothing choke directly following the h.t. rectifier).

The boosted h.t. line is obtained from the dual-standard stage in exactly the same way as it is derived in single-standard circuits. In Fig. 5.5 C_3 is the booster reservoir capacitor, and the boosted h.t. line feeds the picture tube first anode and focusing electrode. The tube final anode is, of course, fed from the output of the e.h.t. rectifier valve.

Field Correction

The boosted h.t. line also powers the field oscillator on many models. Such a feed is shown on the circuit in Fig. 5.2. Here S_1 is concerned, apart from changing the time constant capacitor, with reducing the boosted voltage applied to the oscillator. On 405 lines the boosted h.t. is applied to the field oscillator (top of height control) *via* R_1. On 625 lines R_2 is switched in to chassis, *via* S_1, thereby giving a potential-divider effect, with the field h.t. feed taken from the junction of the divider. The field voltage is thus reduced on 625 lines to compensate for the rise in boost voltage on that standard.

The circuit in Fig. 5.5 also features a stabilizing arrangement, which is also found in some 405-line-only models. It works as follows.

Line Stabilizing

The voltage dependent resistor (VDR) rectifies pulse voltage fed to it from the scanning coils through C_4. Rectification occurs because of the non-linear nature of the VDR. A d.c. voltage is thus developed across

R_2/R_3, and this is applied to the control grid of the line output valve through R_4.

The effect is then similar to an a.g.c. system. Should the pulse voltage amplitude tend to fall for some reason, the negative voltage across R_2/R_3 would also fall. This would then make the control grid of the line output valve less negative, thereby increasing the effective gain of the stage. The stage would operate at greater efficiency and the original drop in pulse amplitude would be automatically compensated.

The operation condition of the line output valve is set initially by the preset width control which can be used to adjust the negative voltage applied to the grid by the required degree of counteracting positive potential picked up from the h.t. line.

Barkhausen Oscillation

Inductors L_2 and L_3 in series with the anode of the line output valve and the cathode of the booster diode respectively are for quelling Barkhausen oscillation. This is a form of electron oscillation which takes place within the line output valve itself, often as the result of operating the valve "below the knee" of the characteristic curve. Below-the-knee operation, however, is not essential when a stabilized line output stage is used, but Barkhausen oscillation may still develop, as also may booster diode switching transients.

The small inductors mentioned combat the effects, which often manifest themselves on receivers not adequately suppressed (or as the result of a faulty line output valve) as watery, vertical lines towards the left of the screen, and also as "windscreen wiper interference" between receivers working on different channels.

There are two main theoretical modes of dual-standard line timebase operation. One is called "constant flyback *time*" and the other "constant flyback *ratio*". In practice, however, a mixture of the two modes is often adopted as, for example, in the circuit of Fig. 5.5.

With a true constant flyback *time* system the *time* occupied by the line flyback is arranged to be the same on both standards, while with a true constant flyback *ratio* system the *ratio* of the time of the scanning stroke to the flyback stroke is arranged to be the same on both standards.

This may, on the face of it, appear to have little bearing on the fundamental operation of the line output stage; but in practice the reverse is true.

If a 405-line line scan and flyback cycle is adjusted to suit the 625-line conditions, both the line scan and the flyback will occur in periods of time about 30 per cent shorter than in the 405-line case. This represents the constant flyback ratio system (the ratio remaining the same since *both* the scanning time *and* the flyback time are reduced by the same amount on changing from 405 to 625 lines). A big factor here is the stepping up of the flyback, since the faster the flyback, the greater the peak voltages produced in the line output stage.

Constant Flyback Ratio

Thus, the constant flyback ratio system yields greater peak voltages which are none too easy to control in terms of insulation. Moreover,

owing to the change in time of the flyback on standard change, the third harmonic tuning of the line output transformer needs to be altered, as we have already seen.

The constant flyback time system, on the other hand, does not increase the peak voltages on 625 lines relative to 405 lines. Also, since the flyback time does not change over the two standards, retuning the loss inductance of the line output transformer to the third harmonic of the flyback is unnecessary. Just one tuned frequency is satisfactory on true constant flyback time systems.

However, *since the ratio of scanning to flyback strokes alters on* constant flyback time systems, compensation is required to equalize the heater current of the e.h.t. rectifier valve over the two standards, again as we have already seen.

Constant Flyback Time

There is another point so far as the constant flyback time system is concerned, and that is since the flyback is usually slowed down on 625 lines to make it the same as that on 405 lines it is possible that the picture signal may start a few microseconds before the flyback has finished. The effect could then be a slight foldover on the left of the picture.

This trouble can be overcome by *making the 405-line flyback a little* faster than normal so that the 625-line flyback need not be slowed down quite as much, bearing in mind that a few microseconds of tolerance are available on the porches of the line sync. pulses. This, then, gives the system which is a compromise between constant flyback time and constant flyback ratio. With such a system the line output stage may have to be switched for third harmonic tuning and it may also be necessary to compensate for e.h.t. rectifier valve heater current. Indeed, with some compromise systems it is often difficult to know whether constant flyback time or constant flyback ratio predominates.

With a flywheel controlled line oscillator it is not unduly difficult to "phase" a picture *within* the raster so as to avoid cutting of the left-hand edge of the picture owing to the picture signal starting a microsecond or so before the finish of the 625 flyback. The picture on both standards can be "phased" and the separate line lock controls set so that the picture occurs in the centre of the raster on both standards. Note that a displaced raster, as distinct from picture displacement within the raster, would be caused by incorrect adjustment of the shift control on the tube neck.

The same effect is accomplished by the use of a switched delay network in the feed between the sync. separator valve and the line oscillator. At least one manufacturer has adopted this arrangement (early dual-standard Ultra) whereby the 625 sync. pulses are delayed relative to the 405 pulses.

Note that with a constant flyback *ratio*, systems may incorporate a small value resistor (typically 2 ohms) in series with the e.h.t. rectifier valve heater as a stabilizing device; but the heater current is held within the tolerance of the valve over the two systems, thereby avoiding the necessity of heater current control.

6

CONVERTING FOR DUAL-STANDARD WORKING

For several years prior to the commencement of the 625-line BBC2 programmes from London, manufacturers were producing so-called "convertible" receivers. At the time of purchase these were suitable only for the reception of the 405-line transmissions.

Their designs were such, however, that conversion to dual-standard working was possible (though not always practical or economic) by varying amounts of internal refitment and adjustment, depending upon the make, type and vintage of the receiver concerned.

Some of the very first convertible models featured no more above an ordinary 405-line-only model than an extra control knob and switch labelled "405-625". On some models this was really a gimmick switch of the age! Sometimes the switch was wired internally and sometimes it was not. When wired it often served to change the line timebase from one speed to the other.

It was sometimes locked to avoid damage to the line timebase in the event of it being turned without certain circuit additions.

At the same time and a little later, models which were more truly convertible were launched. These had plug-in i.f. and signal strips, the design being clearly to allow later conversion.

Eventually, true dual-standard models with and without the u.h.f. tuner were made available, and these are the models which are chiefly being bought today.

As 625-line u.h.f. transmissions become available throughout the country, area by area, viewers with claimed convertible models will want them converted for dual-standard working at the smallest possible cost.

It is not possible within the compass of this chapter to provide details of the conversion method adopted by specific manufacturers. Details of that kind are published by the set makers themselves for limited distribution among their agents and to professional service engineers. In view of the considerable differences in detail between the various conversions it is highly desirable to acquire as much information as possible on a model with which one may not be fully conversant before contemplating the conversion.

In the light of recent experience, the plans originally evolved by the manufacturers for the conversion of early "convertible to switchable" models may be found to be considerably modified. Indeed, it may be discovered that some makers have abandoned their original ideas of conversion altogether and instead provide a complete dual-standard chassis (or chassis section) to replace the original chassis.

The cost to convert some of the very first models can be prohibitive, and since they may have been in use over several years on the 405-line standard, pending the launching of the new programmes, it may well be found more economical, and often technically desirable, to encourage a

viewer with such a set to trade it in for a true dual-standard model when the new programmes eventually reach his area.

Of course, if the set is only a few months old, then something must be done with regard to conversion. In this respect it must be stressed that manufacturers have made great efforts to resolve the technical problems of conversion of both early and recent models, and in the majority of cases full support is forthcoming from such sources.

While conversions differ in detail between make and model, the broad general principles are fairly consistent. True dual-standard models with the u.h.f. tuner fitted are, of course, immediately suitable for the reception of the new 625 programmes. Similar models without the u.h.f. tuner simply require the installation of the correct u.h.f. tuner. Mostly, this involves fitting the tuner to the cabinet, a suitable cut-out being provided, and wiring it into the chassis or plugging it into a socket, depending upon the type and make of the set.

U.H.F. Aerial an extra

A separate u.h.f. aerial socket is available—in addition to the v.h.f. aerial socket—for plugging in the u.h.f. aerial lead, this being necessary in all areas before the 625-line programmes can be satisfactorily received.

Combining filters, for combining the v.h.f. and u.h.f. signals at the aerials to a common downlead, are available, but are suitable only in areas of high signal field where a few decibels of signal loss can be tolerated. In any event, it is usually necessary to separate the signals again at the set end of the downlead so that they can be applied to their appropriate sockets.

Ordinary 405-line-only sets are not convertible to dual-standard or 625-line-only operation, except on an experimental basis, and by knowledgeable persons.

This, then, leaves only the "gimmick-less" convertible to switchable models. Many thousands of these sets were sold (and are being sold still in areas not yet ready for 625 programmes) and almost the same number will eventually require converting.

Most of the genuine convertible to switchable models come in two distinct types, some incorporating (after conversion) two i.f. strips and others adopting the true dual-standard arrangement of a single, switchable i.f. strip.

Survey of Convertible Models

Some Ekco models, for example, are converted by the installation of a transistorized 625-line-only strip. This is used instead of replacing the existing 405-line strip by a new, dual-standard strip. The 625-line strip incorporates the i.f. channel and f.m. sound detector.

The strip feeds the video stage direct and it receives its signal from a push-button u.h.f. tuner. It is fitted alongside the existing panel.

A conversion strip complete with u.h.f. tuner is shown in Fig. 6.1. This relates to Ultra Model 1984C, which was produced before Ultra became part of Thorn Electrical Industries. For Ultra Model 6604 an external plinth is used to house the conversion items.

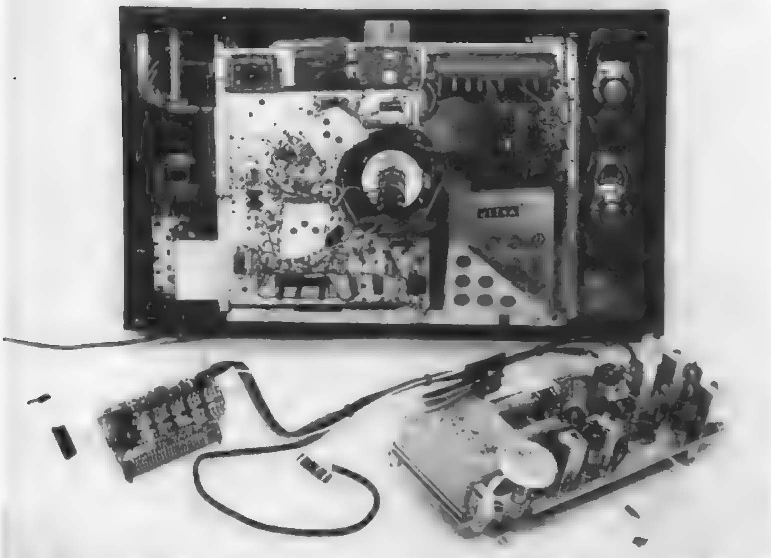


FIG. 6.1. SHOWING THE PLUG-IN CONVERSION ITEMS OF THE ULTRA BERMUDA, MODEL 1984

The "100" series receivers of Bush require the addition of a standard change switch, tapped line output transformer, dual-standard i.f. strip as a replacement for the single standard strip fitted and, of course, a u.h.f. tuner unit.

Models of the "110" series require the addition of a dual-standard i.f. strip substituted for the single-standard strip fitted, plus a u.h.f. tuner.

The more recent TV115L and TV118L models simply require the addition of a u.h.f. tuner, as all the other dual-standard items are fitted. The same applies to the "120" series.

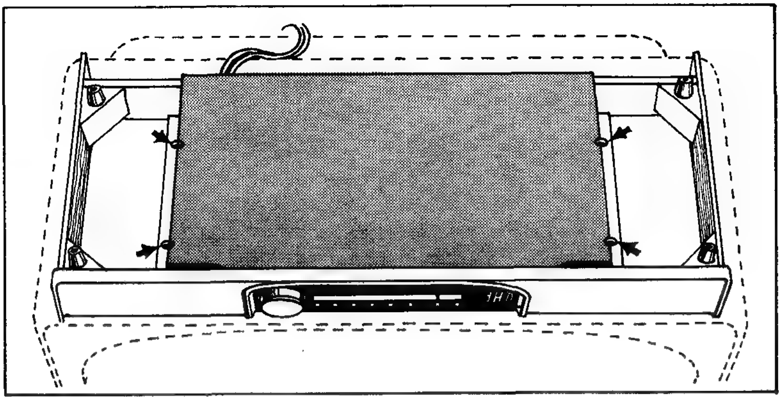
Convertible models of the Murphy range (*e.g.* 659X-689X models and the "700" series) are changed to dual-standard working by external plug-in units and minor alterations to the mains unit. Models of the "800" series simply call for the mounting and plugging in of a u.h.f. tuner.

Baird models of the "600" series have the i.f. strip and u.h.f. tuner as internal fitments with the exception of Model 602, which uses a plinth-type cabinet and screws to the underside of the television set.

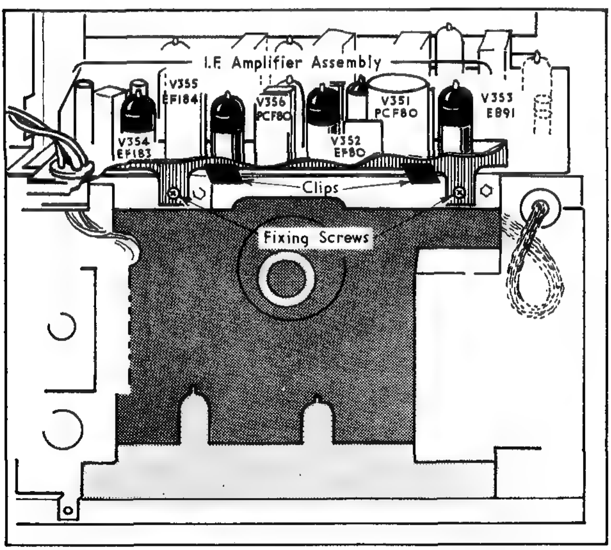
Ferguson "convertibles" use an extra i.f. strip secured to the main chassis and have the u.h.f. tuner fitted inside the cabinet with the exception of Model 3600, which employs a plinth-mounted u.h.f. tuner.

At (a) in Fig. 6.2 is shown how the plinth is installed and at (b) the position and fixing method of the i.f. amplifier assembly. Various ways, depending upon model, of fitting the u.h.f. tuner are shown in Fig. 6.3.

Convertibles of the Philips group of companies (including Cossor, Stella and Peto Scott) are made suitable for dual-standard working by a



(a)



(b)

FIG. 6.2. THE FERGUSON PLINTH U.H.F. TUNER AT (a) AND THE I.F. ASSEMBLY AT (b)

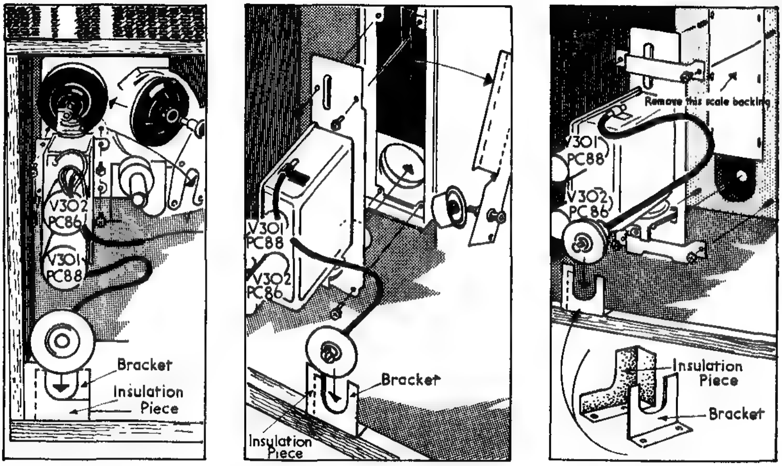


FIG. 6.3. SHOWING HOW THE U.H.F. TUNER IS FITTED IN VARIOUS FERGUSON MODELS

kit of parts containing the u.h.f. tuner, the dual-standard change-over switch, the i.f. strip and a line flywheel unit. These are fitted inside the cabinet and plug into the parent chassis, avoiding soldered connections. The i.f. strip and the u.h.f. tuner of the 19″ models are shown in Fig. 6.4.

In GEC models the existing 405 i.f. panel is removed and replaced with a new 625 version which incorporates the f.m. detector. Valves are taken from the old strip and put into the new one. Two extra valves are required: the new EH90 and an EF183.

On some converted models independent r.f./i.f. circuits are used right up to the video amplifier, coupling being by cathode-follower. Switching standard is by changing over the h.t. supply from one strip to the other, thereby avoiding switching at r.f. This idea is used in some of the Philips group models.

E.H.T. rectifier heater current in this range is kept constant by the switching of a capacitor coupled to the line output transformer by a ferroxcube component. Flywheel sync. is switched in *only* on 625 lines.

On some very early "convertible" models the manufacturers are suggesting that the sets be returned to their service departments for conversion. This may well be worth while in certain cases, for quite a lot of conversion work may be demanded. The sets would also be brought up to full specification.

Although all makes and models have not been mentioned in the foregoing survey, the general idea of conversion methods adopted by most of the makers has been given, bearing in mind that the same basic chassis and conversion procedure are often used in several receivers of different make and model but belonging to the common company group.

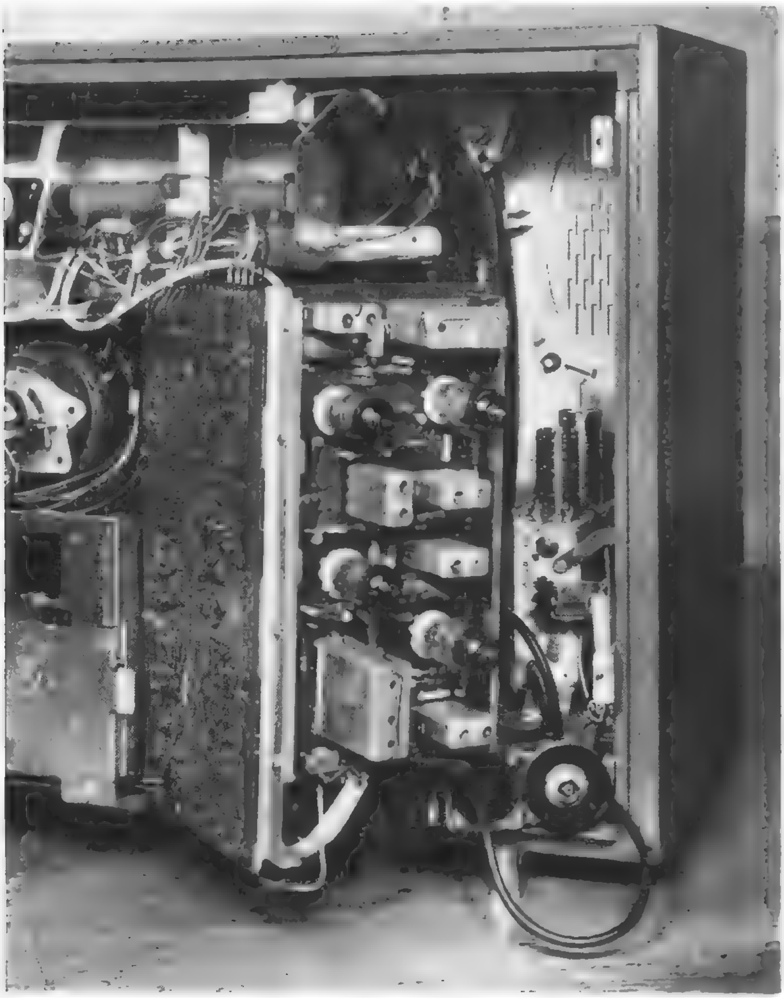


FIG. 6.4. THE PLUG-IN I.F. STRIP AND U.H.F. TUNER OF THE COSSOR RANGE

7

U.H.F. PROPAGATION AND AERIALS

The service area of a u.h.f. transmitter is considerably less than that of a v.h.f. transmitter, assuming equal effective radiated power, aerial height and siting. One reason for this is that u.h.f. signals tend to follow a somewhat less curved path than v.h.f. signals. It is a well-known fact that reasonably good reception of the v.h.f. channels is possible at some distance beyond the line of sight range due to diffraction of the signal around the curved surface of the earth. There is less diffraction at u.h.f. and thus the reception range is less (Fig. 7.1).

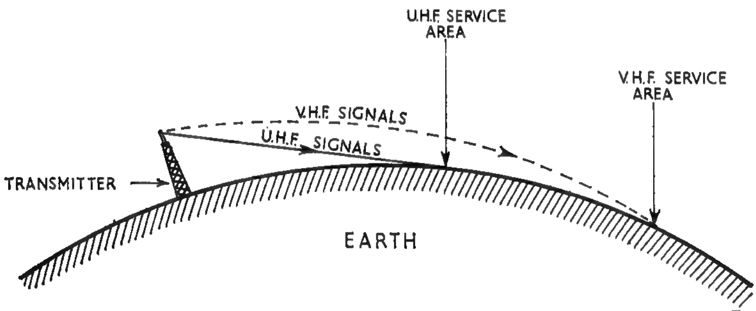


FIG. 7.1. AT U.H.F. THE SIGNALS ARE LIMITED ALMOST BY THE HORIZON AS THEY UNDERGO VERY LITTLE BENDING. AT V.H.F, HOWEVER, THE SIGNALS TEND TO FOLLOW THE CURVATURE OF THE EARTH FOR A SHORT DISTANCE, THEREBY GIVING A GREATER FRINGE AREA THAN AT U.H.F.

Penetration and Diffraction

Moreover, at medium frequencies, signals pass easily through solid objects. At low v.h.f. the signals have more difficulty in passing, but they do get through. As the frequency is further increased, so the signals have even greater trouble in passing through objects such as hills, houses, trees and so on.

At frequencies corresponding to light radiation, even very thin objects are opaque. That is, they prevent penetration of the radiation and result in a distinct shadow area. They may also reflect the light.

While u.h.f. signals are still far below the frequency of light, they do nevertheless have some of the characteristics of this radiation. For example, they do not easily pass through solid objects; they are easily reflected; and they are easily bent or diffracted.

Thus, compared with v.h.f. signals, u.h.f. signals are much more severely attenuated due to solid objects. In practice, this often means that the u.h.f. aerial will need to be mounted as high as possible and well clear of obstacles that could hinder the passage of the signal.

While it is possible to employ attic and indoor aerials on the v.h.f. channels, only in areas very close to a transmitter is it generally possible

to do likewise on the u.h.f. channels. A wet roof, for example, can attenuate the u.h.f. signal by as much as ten times, while only about two or three times loss will occur on the v.h.f. channels under similar conditions.

Reflection

Signal reflections occur at v.h.f. as well as at u.h.f., as is well known. From the practical aspect, any object which is greater than half a wavelength of the signal can act as a reflector. As the frequency is increased, so the wavelength is decreased and at u.h.f. there are very many more objects which are greater than a half wavelength than there are at v.h.f. The potential for signal reflection at u.h.f. is thus far greater than at v.h.f.

Curiously, the reflection at u.h.f. is not as bad as may first be thought. One reason for this is that very high gain u.h.f. aerials are required to make good the propagation and set losses, and as the aerial gain is increased so its beam is narrowed. Thus, signals arriving off beam due to reflection are rejected by the aerial and fail to give the symptoms of reflection, or "multipath interference", as it is often termed.

Another reason is that reflected signals are swiftly attenuated themselves, as are the direct signals, so that by the time they arrive at the receiving aerial, even though the aerial may accept them, they may not be strong enough to cause much trouble.

Nevertheless, trouble can arise due to reflected signals. The signals arrive a fraction of a second after the direct signals, so that to the receiver it "looks" as though two signals of exactly the same transmission are received with a diminutive interval of time between them. The set thus displays the picture due to the direct signal and also endeavours to display the same picture due to the reflected signal. A picture displaced in time to the right of the main picture thus occurs, often called a "ghost picture" or "ghost". The time displacement can be used as a measure to compute the **extra distance** travelled by the reflected signal with respect to the direct signal, should this information be considered warranted or useful.

The phase of the reflected signal with respect to the direct signal can be reversed. The ghost picture is then a negative replica, though weaker, of the direct picture.

Reflections with a very small time displacement can result in the ghost appearing almost coincidental with the direct picture. The ghost may not then be revealed as such, but the overall definition of the main picture is somewhat impaired, depending upon the strength of the reflected signal or signals. Ghosting of this nature can prove extremely troublesome on the u.h.f. bands.

Tests at u.h.f. carried out in Manhattan showed that in heavily built-up areas the picture can appear very blurred—sometimes to an extent equivalent to 0·5Mc/s bandwidth! This has also been experienced in Great Britain as the result of multiple reflections.

The amount of signal induced into a receiving aerial is related to the physical size of the aerial: the larger the aerial, the greater the signal pickup.

However, television aerials need to be tuned devices. That is, their element lengths must bear a relationship to the wavelength of the channel. The highest channel in Band V (Channel 68) has a wavelength

of 0·35 metre. This is approximately 14 inches. For optimum results, however, a television dipole needs to be in the region of half a wavelength which, on the channel being considered, is only about 7 inches. In practice, the dipole is not exactly a half wavelength long, but is in the order of 95 per cent of the theoretical length.

This is because of the physical make-up of the aerial, the inclusion of an insulator and supports for securing the system to a wall or chimney-stack, and the proximity of the aerial to the trees, buildings and so on. The diameter of the conductor comprising the dipole also has a bearing on its length (and bandwidth) and the slowing up of the signal when it arrives at the dipole from free space also comes into the equation.

A dipole tuned to, say, 200Mc/s in Band III has a length of about 28 inches, whereas a dipole tuned to Channel 68 in Band V has a length of about 7 inches. A Band I dipole, at the other extreme, has a length of up to 130 inches.

There are many areas in Great Britain where just a single dipole picks up sufficient signal on the Band I channels to work a television set. At the same location and from a co-sited, almost equal power Band III station, a three-element array is often required to pick up a signal of the same sort of strength.

It is interesting to note that the three Band III elements added together give a total length of similar order to that of a single Band I dipole and, generally speaking, this is why three Band III elements are needed against one Band I element for equal signal induction.

There are, of course, other factors involved, but the foregoing serves to highlight the basic conception.

Signal Grip

The strength of the signal induced into an aerial is related to the amount of metal available to "capture" the signal. To some extent this shows in practice, but the fact should be remembered that, apart from being a tuned device, an aerial must also be designed to work into a specific impedance over a given range of frequencies.

When an aerial is made larger than a simple tuned dipole, it must still tune over the required channel or channels and still "look" like a reasonable impedance match to the feeder.

Generally speaking, a u.h.f. aerial needs just as much, if not more, metal as the aerials on Bands I and III. This extra metal is given to an aerial by directors and reflectors and by stacking two or more complete arrays.

U.H.F. stations often have a greater effective radiated power than their v.h.f. partners. This greater power does not extend the service area by any marked degree, but it does tend to step up the field strength in the reception area, if only by a small amount (by the square root of the power increase).

At this stage it is best to start at the receiver and work back towards the aerial. In the first place, the quality of a picture a receiver is capable of providing is dependent upon the sensitivity and the noise factor of the set.

Tuned to v.h.f. channels, good pictures can normally be obtained when the signal applied to the set is not less than about 100 microvolts. However, on the u.h.f. channels, particularly those on Band V, the same sort of picture requires a signal of at least 350 microvolts. This is because the noise factor of a set tuned to the u.h.f. channels is poorer than when tuned to the v.h.f. channels.

Signal/Noise Ratio

For completely noiseless (without grain) pictures the signal applied to the set from the aerial needs to be about 100 times (40dB) stronger than the noise signal produced by the set itself. Noise signals, it will be remembered, are produced by all amplifiers, and only by adopting very special techniques is it possible to amplify radio signals without the production of significant noise signals.

Thus, if the noise produced by the first stages of a set is equivalent to (say) a signal of 2 microvolts, then for grain less pictures the signal applied from the aerial will have to be, at least, 200 microvolts. This represents a 40dB (100-to-1) signal/noise ratio.

In practice the pictures are perfectly acceptable at signal/noise ratios down to about 20dB. Much below that value the resulting picture grain makes viewing difficult and the definition poor. This would mean that a just-about-viewable picture would be obtained with an input signal of 20 microvolts into a set producing a noise signal equivalent to 2 microvolts (10:1 ratio).

With valve circuits, the frequency of amplification is increased, so the equivalent noise signal rises (the opposite is true with u.h.f. circuits). On the u.h.f. channels the noise produced by the tuner is some two to three times as great as produced on the v.h.f. channels.

This means that to provide pictures of equal grain make-up on, say, BBC1 and BBC2, the BBC2 signal should at least be twice as strong (as applied to the set) as the BBC1 signal.

This adds up to the fact that, in spite of the greater atmospheric attenuation on the u.h.f. channels and the smaller "signal capture", u.h.f. aerials are expected to provide a greater signal voltage in marginal reception areas than the v.h.f. aerials.

There is another factor which aggravates the problem and that is the feeder loss at u.h.f. is approximately four times greater than at v.h.f.

It is most important, therefore, that the pick-up efficiency of the u.h.f. aerial system is at all times kept as high as possible. It is far better to have too much signal on u.h.f. than too little. The healthy development of the u.h.f. system throughout the country depends very much upon the quality of the BBC2 pictures as seen by viewers, and only good and efficient u.h.f. aerials can provide good quality pictures and do full justice to the greater definition potential of the 625-line standard.

Aerial Type

Fortunately, quite complex, multi-element arrays of relatively small overall size are possible on the u.h.f. channels. Most arrays for the u.h.f. bands use a single reflector system and up to about 16 (or more) directors. For greater gain and pickup efficiency two or four complete arrays are

often stacked and correctly "phased" by coaxial transformers and critical spacing.

The reflector system may take the form of a number of elements (such as Antiference) or a toast-rack arrangement (Belling and Lee). A typical u.h.f. aerial by Belling and Lee is depicted in Fig. 7.2. This employs a

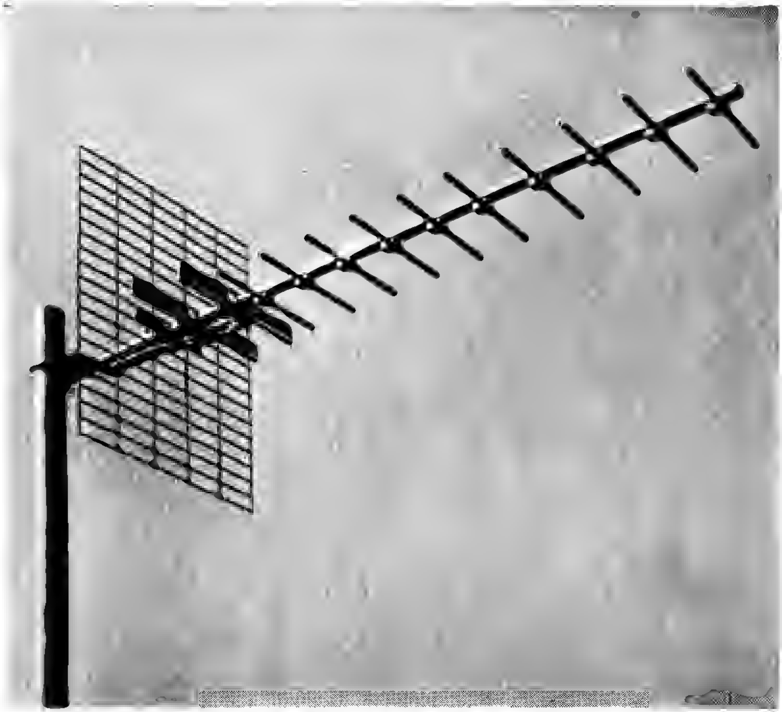


FIG. 7.2. A U.H.F. AERIAL BY BELLING AND LEE. NOTE THE METHOD OF CLAMPING TO THE TOP OF A METAL POLE. IN SOME CASES THE V.H.F. AERIALS MAY BE MOUNTED ON THE SAME POLE BELOW THE U.H.F. AERIAL

pair of "phased" dipoles, a toast-rack reflector and ten directors. An aerial of this kind would have a gain of around 9dB. By the stacking of two such arrays the gain would rise to about 12dB. Of course, as the gain goes up so does the directivity, and too great a directivity can prove an embarrassment owing to highly critical orientation. If the arrays are beamed too much, movement due to wind can cause flutter and fading troubles. This applies particularly to aerials such as the paraboloid and corner reflector, as distinct from the basic Yagi array considered.

It is worth noting that, although a reflector system may be just a wire mesh, or may include a number of elements, it is classified as a single element by aerial manufacturers in Great Britain.

In areas very close to a transmitter one could be tempted into using a very simple aerial, but this is not always worth while since, unless a

reasonable degree of directivity is available, the elimination of forms of r.f. interference would be impossible. There is the possibility of such interference from the local oscillators of receivers tuned to v.h.f. channels, depending upon the harmonic relationship. It is just as well to have some directivity to discriminate against interference of that kind.

Set-top aerials will work in some cases, but proximity effects will tend to spoil the reception and fading may occur as people move about the room. Impedance changes and similar disturbances which result when an aerial is sited in a zone of standing-wave pattern can also have a serious effect on picture definition.

Attic type aerials are quite feasible if the roof attenuation can be tolerated, bearing in mind that the attenuation ratio rises with rain. It is as well to use a complex array in the attic to make up for the signal loss; but keep the aerial well clear of metal fitments and water pipes.

In many cases chimney-mounted aerials will present something of a problem, especially where a shared chimney-stack is already carrying two multi-element Band I arrays, two Band III arrays and, possibly, two Band II arrays. There just will not be enough room to cater for two more Band IV/V arrays! Where there are only one or two aerials on a chimney-stack, the u.h.f. array can be mounted on a pole extension of the Band I aerial. It is always desirable to arrange for the u.h.f. aerial to be above the v.h.f. aerials.

Boost V.H.F. and allow Indoor Aerials

Where the chimney is already cluttered some other line must be taken. Where outside u.h.f. arrays are essential for good pictures, the v.h.f. aerials could be attic-mounted and, if necessary, a small low-noise dual-band v.h.f. preamplifier could be used to step up the signal which will by this action be reduced. Transistorized dual-band amplifiers of this kind are available. Such an amplifier, either battery or mains powered, could be used to supply the set of an adjoining neighbour from one aerial system.

Bringing the v.h.f. aerials indoors by this means will permit the u.h.f. aerials to take advantage of the chimney mounting position. It must be stressed that height and free space are essential for the u.h.f. aerial.

Up to seven directors are often needed for u.h.f. reception within a radius of 15 miles; eleven directors up to 20 miles; sixteen directors up to 30 miles; and stacked arrays for fringe areas beyond 25 to 30 miles, and in difficult locations nearer to the transmitter where extra gain and directivity are required.

In difficult and fringe areas special u.h.f. feeder cable should always be used. If in doubt, use cable with an inner conductor of not less than 0·044 inch diameter.

Some aerials for the u.h.f. channels have a so-called "balun" transformer and others have not. A "balun", as its name implies, is a device which makes a balanced circuit unbalanced in one direction and an unbalanced circuit balanced in the other direction.

A dipole is balanced while coaxial cable is unbalanced. Theoretically, therefore, a balun should be employed to connect the aerial to the feeder, close to the dipole. In practice, especially with v.h.f. aerials, this technique

is rarely used. Owing to the more critical nature of the u.h.f. signals, however, some authorities claim that a balun is vital while others say that it is still not essential.

Some u.h.f. aerials, therefore, will be found with a balun and others without.

Transistorized boosters for set and aerial mounting are also available to increase the gain of the aerial system and to improve the signal/noise ratio of the installation.

Matching

A factor of major importance is matching between the aerial and feeder. The centre impedance of a dipole is about 80 ohms. The addition of directors and a reflector reduces this value. Folding the dipole raises the dipole impedance again. A simple fold increases the impedance by about four times.

This folding technique is adopted on all multi-element arrays to facilitate matching to the feeder. With u.h.f. aerials, though, a simple fold may not be the complete solution and some aerials utilize a dipole of inconsistent length/cross-section ratio.

Bandwidth

By virtue of its ultra high frequency design, a u.h.f. aerial usually has a bandwidth in excess of that for v.h.f. aerials. An aerial on Band I, for instance, is suitable only for one specific channel. The same may apply to an aerial on Band III, but, here, a bandwidth embracing two or three channels is possible.

It was explained in Chapter 1 that there will be four ultimate u.h.f. channels in any reception area which, with guard channels to separate them, will embrace a bandwidth equal to eleven 8Mc/s channels (*e.g.* 88Mc/s).

U.H.F. aerials are thus being made responsive over such a bandwidth, so that **one** correctly dimensioned aerial will embrace all four channels. Note, however, that in certain areas it may be necessary to depart from the 88Mc/s principle. In those cases more than one u.h.f. aerial could eventually be required to obtain the four programmes, but aerials arranged to respond over the whole of the u.h.f. bands will become available in time.

8

SERVICING DUAL-STANDARD RECEIVERS

THERE are no fundamental differences between the procedures adopted for the servicing of single and dual-standard sets. A technician familiar with the servicing techniques of 405-line-only receivers will accommodate dual-standard models in his stride, once he is acquainted with a few items of detail. Indeed, there will be times when the dual-standard feature will actually assist with the fault diagnosis.

Take for example a receiver which produces e.h.t. voltage on 625 lines but not on 405 lines. The technician will know immediately that the line timebase valves are good. Moreover, since the line output transformer is also common to both standards, the symptom would mostly signify the goodness of that component, as well as others which are common to both line speeds.

A set exhibiting the symptom would probably have trouble in the time constant (including the 405-line hold control) components of the line oscillator or in the associated switch section or sections.

It should be noted at the outset that dual-standard sets do give trouble in their "standard change" switch sections.

Slider Switch

Most models employ a "slider type" switch extending the length of the chassis. The switch is designed in conjunction with the mechanics of the chassis or printed circuit board so that the individual sections of the switch are placed as close as possible to the circuits which require switching

We have seen in previous chapters that many circuits must be switched to change their conditions to suit the needs of the selected standard and to ensure that the circuits operate at maximum efficiency on both standards. In the line timebase fairly high amplitude pulse voltages are switched. High resistance switch contacts could thus result in sparking or arcing across the contacts. Initially, this may simply produce a few flashes across the picture, but eventually the contacts are caused to burn and become soft, thereby destroying the switch action completely or considerably impairing the operation of the circuit.

Circuit switching in the signal stages suffers similar faults as those eventually experienced on wavechange switches of ordinary radio sets. Fortunately, standard change switches in television sets are much more robust than wavechange switches in radio sets, so it is usually only after a period of considerable use that normal switch wear begins to show.

Insulation breakdown, though rare, can happen between switch sections. In such a case the negative bias on a grid circuit may gradually become neutralized by a positive "leakage" potential.

The slider of the standard change switch is mechanically coupled to the v.h.f. tuner, so that in the U.H.F. position the switch is automatically set for 625-line operation. This technique is made possible, of course, because the 625-line programmes are, at the time of writing, transmitted only in the u.h.f. bands in Great Britain.

In Eire the v.h.f. bands are used for both 405 and 625 lines, and slightly modified receivers without u.h.f. tuners are available. The same type of receiver is coming into use on coaxial relay systems in Great Britain, for at the master aerial stations of such systems the u.h.f. signals are translated to v.h.f. signals for accommodation in an unused Band I or III channel. The mechanical coupling to the standard change switch is then altered (or some other arrangements are made, depending upon the type of set) so that the receiver goes over to the 625-line standard when the tuner is set to that v.h.f. channel carrying the 625-line programme.

The service technician will discover certain valves in dual-standard models that are not used in single channel models. These have already been referred to in previous chapters, but for the sake of completeness they are detailed below.

Valves for Dual-Standard Sets

Valves designed for u.h.f. operation and tuner service include the two triodes PC86 and PC88. The latter works as an r.f. amplifier and the former as a self-oscillating mixer. Both valves use frame grids and so possess a high value of mutual conductance, with grid lead inductance and internal capacitances kept to a very minimum.

A triode-pentode valve, with each section of frame grid construction, is the PCF801. This is often found working as an oscillator/mixer in v.h.f. tuners and, when switched, as a controlled i.f. amplifier following the u.h.f. tuner.

New valves in the line timebase include the PCF802, the PL500 and the PY88. The first is found in flywheel controlled line oscillators, the pentode acting as a sine wave oscillator and the triode as a reactance valve which itself is controlled by a potential derived from a discriminator circuit.

The PL500 is a line output valve which has an exceptionally high ratio of anode current to screen grid current, achieved by a new form of anode, called the "cavitrap". This is capable of delivering high values of deflection power and readily accommodates the higher energy requirements of the 625 line speed.

The PY88 is a booster diode. It has a heater/cathode rating of 6·6kV and peak and average current ratings of 550 and 220mA respectively. In a stabilized line timebase, the PY88 is often used together with the PL500.

Other dual-standard valves include the noise-cancelling ECH84 sync. separator and the heptode EH90 f.m. detector. At the time of writing, valves employing double-pentode frame grid sections are being considered.

Probably the most critical section of a dual-standard receiver is the u.h.f. tuner. Many authorities recommend that repairs to this unit should not be attempted by the service technician and that a defective or suspect tuner be returned to the manufacturer's service department for specialized attention.

U.H.F. Tuner

While it is true that u.h.f. tuners are extremely critical units, it would seem that most service engineers will be able to undertake normal servicing procedures provided full account is taken of the ultra high frequency circuit techniques adopted.

Valve replacement could put the tuning of a specific channel in a slightly different position on the control from that to which the viewer has become accustomed, owing to capacitances changes. A valve change could probably affect the sensitivity of the tuner to an extent demanding internal adjustments—but this is not very likely. If a valve change really upsets the tuner, the valve itself could be out of tolerance and a second valve should be tried before the tuner is sent away or examined in detail.

When v.h.f. tuners were first introduced they were looked upon with awe and it was the order of those early Band III days to send the tuner back to the maker should it appear faulty. V.H.F. tuners are still returned to the factories, but mostly the service technician himself sets out to clear the trouble and to undertake the final trimming. There is not much point in manufacturers telling service engineers NOT to attempt a repair.

Signals in the u.h.f. channels are up to four times as high in frequency as those in Band III. This means that a u.h.f. tuner is up to four times as critical as a v.h.f. tuner on the Band III channels.

It is known, of course, that bad mistuning can result on the Band III channels merely by forgetting to replace the screening cover at the bottom of the v.h.f. tuner. By forgetting the same operation on a u.h.f. tuner, the result could be total detuning and lack of signals.

It is imperative, therefore, to ensure that all screens are thoroughly secured on u.h.f. tuners, and this applies also to the valve shields. Aside from mistuning troubles, a tuner with a missing or loose screen can also let out signal from its local oscillator. While this may not disturb the operation of its own circuits, it can severely interfere with nearby (and sometimes distant) receivers also working on the u.h.f. channels.

Patterns result on the affected set or sets as the u.h.f. tuning of the radiating set is adjusted.

Poor u.h.f. tuner screening can also cause breakthrough of the BBC1 sound programme on 41·5Mc/s. If the symptom is troublesome normally, however, it should be countered by careful setting of the u.h.f. tuning control.

While on the subject of screening, it should be noted that inadequate screening of the v.h.f. tuner can lead to interference problems on the u.h.f. bands.

In the early days of television, in areas where the BBC programme was obtainable on two Band I channels (*e.g.* the Oxford area, where the BBC programme could be received on Channel 1 London and Channel 4 Sutton Coldfield), local oscillator radiation from sets adjusted to one channel resulted in bad pattern interference on nearby sets taking their signals from the other channel.

Radiation Troubles

This trouble rarely arises nowadays since most of the country is adequately served by carefully spaced and channelled v.h.f. transmitters.

However, harmonics of the v.h.f. local oscillator, when the tuner is adjusted to certain channels, fall within the passband of certain u.h.f. channels and can cause pattern interference.

Following tests conducted jointly by the BREMA and the GPO, recommendations calling for tuner radiation limits of 600μV/metre for harmonics from v.h.f. tuners and 3mV/metre for harmonics from u.h.f. tuners (measured at a distance of 3 metres), are in the hands of set and tuner makers; but, of course, these limits will be greatly exceeded if the screening is inadequate following a servicing operation.

We have seen, of course, that u.h.f. channel allocations have been made to minimize the likelihood of trouble from this cause, but difficulties can arise in some limited areas, especially as booster stations are brought into service. One way of combating pattern interference troubles is by ensuring that the u.h.f. aerial is as efficient as possible.

On the v.h.f. bands, adjustment of the fine tuning control causes the sound and vision carriers to traverse the sound and vision i.f. response curves. This causes a variation in the quality of the picture and an alteration in the level of the sound as the carriers move from their optimum positions. Sound-on-vision interference results when the sound carrier enters the region of the vision response curve, while vision-on-sound is heard when the vision carrier is displaced sufficiently to cause low frequency components of the vision signal to spread into the sound channel. These are the normal effects which occur when the v.h.f. fine tuning control or oscillator is adjusted.

U.H.F. Tuning Effects

The effects differ somewhat on the u.h.f. channels. The intercarrier sound frequency cannot alter since this is fixed at 6Mc/s by the difference between the frequencies of the sound and vision carriers at the transmitter.

The vision carrier runs along the vision i.f. response curve and, as before, this changes the picture quality. However, because the levels of the sound and vision signals arriving at the vision detector change as the u.h.f. tuning is adjusted, so the ratio of levels deviates from that required for optimum intermodulation to produce the intercarrier signal (sound level about 30dB below that of vision), and sound distortion results.

Further, sound distortion is also caused because of the resulting unbalance of the f.m. sound detector, and intercarrier buzz becomes troublesome. Thus, although drift of the u.h.f. tuner local oscillator will not particularly cause sound fade, as at v.h.f., it can cause sound distortion *and* intercarrier buzz.

No Sound without Vision

Another point worthy of note is that no sound signal is produced unless the vision carrier is present at the vision detector. Failure of the vision signal at the transmitter, therefore, also cuts off the sound. This never happens on the v.h.f. 405-line standard, so is something entirely new. Another new feature on 625 lines is the apparent lack of the line whistle. This is clearly heard by most on 405 lines at 10,125c/s, but far fewer viewers and engineers have hearing which extends adequately up to the 15,625c/s line speed of the 625-line system.

Alignment of dual-standard receivers should never be undertaken without reference to the maker's service manual and alignment instructions. While it is often possible to align 405-line-only sets aided only by Test Card C and a bit of knowhow, this technique is virtually impossible on dual-standard models, since many more filters for response shaping are used in the vision i.f. channel than in 405-line-only sets.

Reasonable results may be achieved on 405 lines without instruments and a data sheet, but it is extremely unlikely that results would be at all encouraging on 625 lines. Then, if adjustments are attempted to bring 625 lines in on Test Card C, the performance previously considered reasonable on 405 lines would deteriorate badly.

Intercarrier Sound Alignment

The dual standard intercarrier sound channel is another critical circuit. This must always be centred spot-on 6Mc/s, for the fine tuning control or channel tuning cannot be utilized on the u.h.f. channels to move the sound carrier to the top of the sound response curve, as we have already seen.

In a dual-standard model the intercarrier tuned circuits are comparable to the oscillator tuning in a 405-line-only sound channel. Unlike the oscillator, however, the intercarrier tuning cannot easily be altered. It is thus essential that the alignment here be very carefully carried out in the first place.

A signal generator with crystal calibration is desirable to ensure that the channel is centred on 6Mc/s. If such an instrument is not available, the next best thing is to beat the signal generator against the 6Mc/s in the sound channel, assuming that the set is not completely dead.

If some signal is getting through the intercarrier channel, the generator should be loosely coupled to the amplifier input, the signal level turned up with the modulation off and the tuning adjusted for a beat whistle and consequent dead-beat point.

As a last resort, the intercarrier tuning can be adjusted for maximum voltage across the d.c. load of the ratio detector when the set is actually receiving a signal. The ratio detector coil itself can then be adjusted for optimum balance, which is generally signified by minimum intercarrier buzz.

This is not one of the best ways of making the adjustments, of course, since it is not a good thing to peak the intercarrier circuits, for a certain amount of tolerance should be allowed for circuit drift; but it does, at least, allow a receiver to work better than when the intercarrier and f.m. detector circuits are badly maladjusted.

While there are a number of circuits in dual-standard models which have no direct equivalent in single standard models, the operation of these can hardly be considered as complex in relation to the very many circuits with which the practising service technician is fully conversant and which, anyway, represents the majority of a dual-standard set!

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